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AN X-BAND FREQUENCY MODULATED RELAY SYSTEM FOR VIDEO FREQUENCIES

REPORT

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Report 977 January 3, 1946

AN X-BAND FREQUENCY MODULATED RELAY SYSTEM FOR VIDEO FREQUENCIES

Abstract

Section 1. A description of an X-band system for transmitting video-frequency signals over distances such that several relaying points would be involved. The criterion of performance used was that the received signal should produce as good a PPI presentation as the unrelayed signal. With the two transmitters and two receivers that were built, it was possible to simulate one intermediate relaying point and for this transmission condition no difference could be detected between the two PPI presentations.

Section 2. This section concerns the receiver from the input of the i-f amplifier to the video output. A few general remarks regarding the design of FM receivers are made including noise considerations, gain and band-width requirements and limiter design. This is followed by a more detailed description of this receiver.

Section 3. A description of means by which two channels were to be transmitted and received on single antennas with a proposed frequency allocation chart.

Section 4. Description of the proposed antenna and tower installations.

Lowell M. Hollingsworth
H. Logemann, Jr.
A. W. Lawson, Jr.
J. M. Sturtevant

Approved by:

A. E. Whitford
Leader, Group 53

A. G. Hill
Head, Division 5

Title Page
26 Numbered Pages
1 Unnumbered Page
31 Pages of Figures

AN X-BAND FREQUENCY MODULATED RELAY SYSTEM FOR VIDEO FREQUENCIES

Section 1.

General Considerations

The apparatus described here was actually the third equipment considered for relaying video information. The other equipments were to be employed for other types of service in which non-directional antennas would have to be used. With non-directional antennas, a low frequency gives superior performance. However, when a decision was made to build equipment for only ground, point-to-point service, the logical decision was to construct it for use on the X-band.

The great effect of wavelength upon range may not be appreciated unless one has seen the result of calculations. If it is assumed that parabolic antennas will be used at both transmitter and receiver and that the useful range will be reached when the signal is 40 db above noise, the following results are obtained for a transmitter power of 100 mw:

Frequency mcs	Wavelength cm	Range - miles 9-ft. Parabola	3-ft. Parabola
300	100	8.8	.975
1000	30	29.2	3.25
3000	10	88.0	9.75
10,000	3	292.0	32.5

These figures indicate that a very large antenna operated at a very short wavelength would be capable of extreme ranges. Actually, the height of the antenna soon becomes much more important than antenna size. The line-of-sight distance should be taken as the extreme range which might be employed. For example, with 50-ft. towers, this distance is 20 miles and with 100-ft. towers it is 28 miles. The other point that might be overlooked is that the increase in range which is obtained by large antennas means that energy is being focused into a smaller area. Thus at the X-band, a 9-ft. parabola would have a beam width of 0.8° while a 3-ft. parabola would have a beam width of 2.4°. This latter figure was thought to be more practical for a portable installation in which the antenna mounting might not be as stable as could be desired.

The video signals to be transmitted were those obtained from the MEW output which is 1.5 volts positive. The video response of the MEW is about one megacycle so it was assumed that if the video response in the relay link exceeded this value, satisfactory operation would result.

The low power requirement made practical the use of the 2K39 (419 Klystron) as a transmitter tube. Some thought was given to the possibility of using the cross-band system proposed for communication work[1], but because of the difficulties attendant with getting this

1. RL Report 830, A Duplex Communication System for Microwaves, R. V. Pound, November 20, 1945.

system to operate successfully at video frequencies it did not seem feasible at this time. It was decided that a separate transmitter and receiver would be designed, and that the unmodulated frequency of the transmitter would be stabilised against a cavity. The local oscillator in the receiver would also be stabilised against a cavity thus simplifying the automatic frequency control problem. A method for stabilising these oscillators had been previously developed.[2]

Means of applying modulation to the transmitting tube was different from that previously employed. It was decided that the stabilising amplifier would be dc coupled to the reflector of the Klystron while the modulation would be ac coupled to the same point. This meant that the highest frequency for which frequency stabilization could operate would necessarily be less than the lowest modulation frequency. Some difficulty from this arrangement might be anticipated but actually no trouble was experienced.

At first there was considerable discussion regarding the merits of amplitude and frequency modulation for this type of service. The general conclusion indicated little difference between the performance of the two and frequency modulation was chosen since it could be accomplished with less difficulty. A deviation of six megacycles was talked of at one time and became the accepted value. The band width required for a deviation of six megacycles was also the subject of much discussion as was the question of whether the cross-over point of the discriminator should be located symmetrically with respect to the pass band. With pulse-type modulation, the frequency spectrum is not symmetrical as it would be for sine-wave modulation. Rather there exists a continuous component at f_o, the undeviated frequency, surrounded by a spectrum corresponding to an amplitude-modulated pulse of the same length as that being transmitted. This is actually the time that the transmitted frequency is not at f_o but is at f_d, the deviated frequency. Then at f_d there exists a spectrum corresponding to the same amplitude-modulated pulse. There is interaction between these two spectra leading to a line-type spectrum. The conclusion is that the actual band width required is only greater than the difference between f_o and f_d by the width required for the spectrum of the shortest pulse to be transmitted. Practically, some additional width must be allowed for possible mis-tuning of receiver or transmitter.

It had also been shown that if the discriminator was symmetrically arranged with respect to the pass band, any uniformly distributed noise would balance out. This was believed to be a desirable arrangement under some conditions of service but since it would require such a wide intermediate-frequency amplifier, it was considered impractical in the present set.

Transmitter

The transmitter is shown in Figs. 1-1, 1-2, and 1-3. The top unit is the oscillator power supply; the next unit is the reflector supply;

2. An Improved Frequency Stabilization System for Microwave Oscillators, R. V. Pound, October 26, 1945.

the next panel mounts the transmitter tube (2K39), the modulation amplifier, and the r-f components; and the bottom unit is the oscillator frequency stabiliser. A block-type circuit diagram is shown in Fig. 1-4. Detailed drawings for the three units listed previously are given in Figs. 1-5, 1-6, and 1-7. The circuit of the modulation amplifier which is located in back of the panel near the 2K39 is shown in Fig. 1-8.

A logical point at which to begin an explanation of the transmitter is with the frequency discrimination which is shown about the center of Fig. 1-4. This consists of two magic tees, two crystals, and a resonant cavity. Theory of operation has been discussed elsewhere[2] and will consequently be brief here. Power from the main line enters through a 10 db directional coupler and a variable attenuator. At the first tee, power divides between the two arms, one of which is short circuited and the other is terminated by a resonant cavity. With the cavity tuned to resonance, the position of the shorting plunger is such that the two reflected waves will arrive at the tee 90° out of phase. As a result, half of the power will flow to crystal No. 1 and half will go back to the first tee where it will divide between crystal No. 2 and the input arm in which it is absorbed. If the cavity is off resonance, the phase of the reflected wave from this arm will change so that the phase difference is more or less than 90°. Consequently, more or less power will go to crystal No. 1 and the inverse is true of crystal No. 2. This results in the two crystals having outputs which change in opposite directions near resonance. These are connected to the inputs shown in Fig. 1-7 with crystal No. 1 connected to the input containing the potentiometer. Two stages of dc coupled amplification are employed with some inverse feedback. Because the cathodes are returned to such a large negative voltage, the output must balance irrespective of the input balance condition. The potentiometer in the plate circuit of the first stage is for setting initial balance. Output voltage is taken from one plate through series resistance to a SPDT switch. Two resistances and a potentiometer are connected across the supply voltage and the tap on the potentiometer is connected to the SPDT switch. This makes it possible to have the actual output voltage either from the amplifier or from the fixed source.

As also shown in Fig. 1-4 the control voltage is 200 ± 100 volts and is connected to the positive terminal of the reflector supply, Fig. 1-6. The negative terminal connects to the modulation amplifier and also to the clipper circuit located in the oscillator power supply. A lead from the oscillator power supply brings back the -1200 volts for use on the voltmeter so that cathode-to-reflector voltage may be measured. In the modulation amplifier, the complete circuit of which is given in Fig. 1-8, the negative voltage feeds through a 100K resistor to the reflector lead. It may be noted that the condenser from the plate of the 6AC7 to the reflector of the 2K39 has a voltage rating of 3000 volts. Lower rating condensers were used but broke down in a short time. Care was taken in insulating the case of this condenser to keep the capacitance to ground reasonably low. The oscillator power supply was used for all other voltages applied to the 2K39 except the reflector voltage.

By the means outlined here, it is possible to hold the frequency of the 2K39 to a value determined by the resonant cavity. When properly adjusted, the 2K39 will follow the cavity frequency to the extent of its electronic tuning range. To adjust the 2K39 for operation on a given frequency, it is first tuned in the usual manner on the fixed voltage position available from the stabilizer. The balance control in the stabilizer amplifier has previously been adjusted to bring the balance voltmeter to zero. When operating at the correct frequency, the cavity is tuned well off resonance and the input potentiometer is adjusted to make the stabilizer-balance meter come to zero. It may also be necessary to adjust the input power by means of the attenuator. The cavity should then be tuned to insure that adequate response on both sides of resonance is obtained. After this is satisfactorily adjusted the cavity should be left at resonance and the switch on the stabilizer should be thrown to the other position so that the reflector voltage is determined by the amplifier. It may be necessary to change the voltage output of the reflector supply in order to obtain oscillations. The tube should oscillate at the same reflector voltage as previously so the voltmeter will indicate in which direction this voltage must be changed. The reflector voltage should be adjusted to bring the balance meter to zero and the mechanical tuning should be adjusted so that the 2K39 is operating at the center of the mode.

Transmitter Test Equipment

Any amount of test equipment might conceivably have been suggested for use with the transmitter. One of the most useful pieces of apparatus would have been a spectrum analyzer. One of these is very advantageous in checking both frequency and frequency deviation. The spectrum analyzer was extensively used in developing the equipment but it was hoped that it might be needed in later equipments only in case of serious trouble.

Other test equipment was installed in an auxiliary unit as shown in Fig. 1-9. Here is located a square-wave generator, a video amplifier and a thermistor bridge. Referring to the left portion of Fig. 1-4, the thermistor bridge is shown connected to the thermistor which receives energy from a 30 db wave selector. In the pictures showing the front of the transmitter, it is the lower wave selector which is marked "30 db power monitor". The thermistor is contained in the mounting to which the wire is connected. This equipment allows a continuous monitoring of power output and is useful in verifying that the 2K39 is adjusted to the top of the mode since that is the point at which greatest power output will be obtained.

Just above the wave selector which is used for power monitoring purposes is a second wave selector to which is connected a frequency discriminator. This discriminator is the same as the one used for stabilization except that a low-Q cavity is employed. It is intended that the discriminator curve be sufficiently wide to accommodate the frequency excursions due to modulation. The output of one crystal is fed to a video amplifier which is located at the top of the rack shown in Fig. 1-9 and for which the circuit diagram is given in Fig. 1-10. (The second crystal is not used.) This amplifier produces a video output having the standard 1.5 volt positive peaks. When presented on a suitable synchroscope this constitutes a monitor receiver as noted on Fig. 1-4.

By tuning the cavity well off resonance the frequency discriminator characteristics can be ruined. Any modulation showing on the scope under these conditions is due to amplitude modulation. By adjusting the mechanical tuning adjustment of the 2K39 this modulation may be reduced to a minimum and consequently one knows that the 2K39 is operating at the top of the mode.

In the actual apparatus as shown, there is a wavemeter in the line leading to the monitoring discriminator. With this arrangement of discriminator and wavemeter it was hoped that frequency deviations might be easily and rapidly checked. This required another piece of equipment shown in Fig. 1-9, the square-wave generator. The circuit diagram, Fig. 1-11 and Fig. 1-12, indicates that provision has been made to feed in a trigger from a synchroscope to synchronize the oscillations. The square-wave output would be employed to modulate the 2K39 and when the output of the monitor receiver was observed on the cathode ray tube of the synchroscope a square wave should result. It was hoped that tuning the wavemeter would affect the square wave response so that frequency deviations could easily be read directly from the wavemeter. Insufficient work was done to prove whether or not the desired degree of accuracy is possible. It may be that the rate of rise of the square wave was too rapid and that slowing it down would improve the performance.

Receiver - R-F Components

As previously mentioned, the local oscillator used in the receiver is stabilized in the same manner by which the transmitting tube is stabilized. The complete receiver is shown in block diagram form in Fig. 1-13 and views of the complete receiver are shown in Figs. 1-14, 1-15, and 1-16. Approximately at the center of Fig. 1-13 the 723 (2K25) local oscillator is shown feeding into a magic tee junction at which point power divides half flowing to the left and half to the right. Power flowing to the left is used to stabilize the frequency as previously described. That which flows to the right goes through an attenuator to a magic tee mixer. It should be emphasized that this is a magic tee and that, consequently, while the two crystals receive both local oscillator power and signal power there is no coupling between the two sources. The signal is fed in at the H-plane junction so that signal energy reaches the position of the two crystals in phase. Local oscillator power enters the E-plane junction and consequently reaches the position of the two crystals out of phase. When detected, the difference frequency at the two crystals is out of phase. By reversing one of the crystals, the i-f signal is reversed and the input circuit to the i-f amplifier can essentially parallel these two inputs. The difference in the two crystal holders may be noted in Fig. 1-15. In addition to causing local oscillator power not to be radiated, there is an improvement in noise figure when this mixer is used. The improvement comes from the fact that any noise components present in the local oscillator input are balanced out and the noise figure is improved over previous designs by this amount.

The intermediate frequency amplifier and subsequent details of the receiver are discussed in detail in Section 2 of this report.

Parts Lists

Since it would be almost impossible to build one of these transmitters or receivers without dimensional information about the radio-frequency components, drawing numbers are listed in Table 1-1.

Future Possibilities

This project was terminated suddenly due to the end of the war and several plans that had been made for the future had to be dropped. Among these was, of course, the change from the present wavelength to one near 3.5 cm. Some of the parts listed in Table 1-1 are for this wavelength. The building of cavities for this wavelength which would be sealed, constant-frequency type was contemplated. These would be assigned frequencies as suggested in Section 3 of this report. It was planned that the first cavities would be constructed of Invar although this would not give the stability that would ultimately be desired. If greater stability was required than these cavities were inherently capable of accomplishing, it might be possible to improve their performance by holding temperature constant. A temperature-compensated cavity might later be designed. In either case, it was planned that the cavities would be hermetically sealed. Use of these fixed-tuned cavities would alter the tuning procedure as given in this report but tests indicate that it would be possible to tune the system satisfactorily.

An improved 419 was being developed which would have had a waveguide output constructed as a part of the tube. This would insure better power output over the band since the transition from coaxial line to waveguide would then be part of the tube. Improved power output by other means were also under consideration as was the design of the tube for better performance at 3.5 cm.

Test Performance

A rather elaborate testing program had been outlined previous to the end of the war. It was subsequently decided to simplify this considerably so that the testing would check the frequency response and distortion of the system but not its range. For this purpose, the two transmitters and receivers that had been constructed were taken to Bedford Field. One transmitter and receiver were installed near the MEW installation so that video from the MEW could be fed to this transmitter. The other transmitter and receiver were taken across the field to the RRL hanger. There the receiver output was fed to the transmitter input so that this formed an intermediate relay point. The receiver at the MEW set picked up the distant (?) transmitter and its output was fed to a PPI scope. This scope was placed immediately adjacent to a PPI scope and both were driven directly from the MEW. This was done so that video response might be judged without the difficulty of decoding rotational information by some means. When the two presentations were compared no difference could be noted; any signal that could be distinguished on one scope could be distinguished equally well on the other. A series of simultaneous pictures were taken and three of these are included as Fig. 1-17. In the two lower series of pictures, the tracks of planes are plainly discerned since the exposures were made for four complete rotations.

Conclusions

A relay system has been designed and tested which appears capable of accomplishing the type of performance expected of it. The video response is so good with one intermediate relay point that it appears probable that two additional relay points could be employed without serious deterioration of the reproduced image.

Lowell M. Hollingsworth
November 15, 1945

Table 1-1

X-Band Relay Transmitter Drawings

Dwg. No.	Component
A-14838-A	Block Diagram of MEW Relay
	Oscillator Power Supply
	Reflector Supply
	Oscillator Frequency Stabilizer
	Modulation Amplifier
	Video Amplifier for Monitor Receiver
	Square Wave Generator for Calibrating Frequency Deviation
A-6991-A	TBN-3EV Thermistor Bridge for Power Monitoring
	Report M-186A
C-6267-A	Type N to Waveguide Flange TPX-44GM
C-12275	10 db Directional Coupler for Frequency Stabilizer
C-14868	30 db Double Directional Coupler for Power Monitor and Monitor Receiver
C-5023	H-Plane Bend in Waveguide
D-13242	Magic Tee R-F Discriminator 3.2 cm
C-9973-A	TPX-34GM Flap Attenuator
B-14880-A	TPX-390A Frequency Meter for Monitor Receiver
A-11653-A	TPX-30EC Frequency Meter for Stabilizer
B-14872	Outline Drawing of 419B Oscillator
B-14873-A	Wavemeter for MEW Relay

X-Band Relay Receiver Drawings

T-6367	Local Oscillator Frequency Stabilizer
D-13242	Magic Tee Discriminator for Frequency Stabilizer
D-12021	Balanced Mixer, Single Magic Tee with Attenuator
	Receiver, Frequency Modulation, 60 mc I-F
B-14886	Block Diagram MEW Relay Receiver
	Drawings Covering Proposed Change in Wavelength
B-14273	17 db Directional Coupler 3.5 cm
B-14274	30 db Directional Coupler 3.5 cm
B-14521	Magic Tee Coupler 3.5 cm
D-14593	Magic Tee Discriminator 3.5 cm

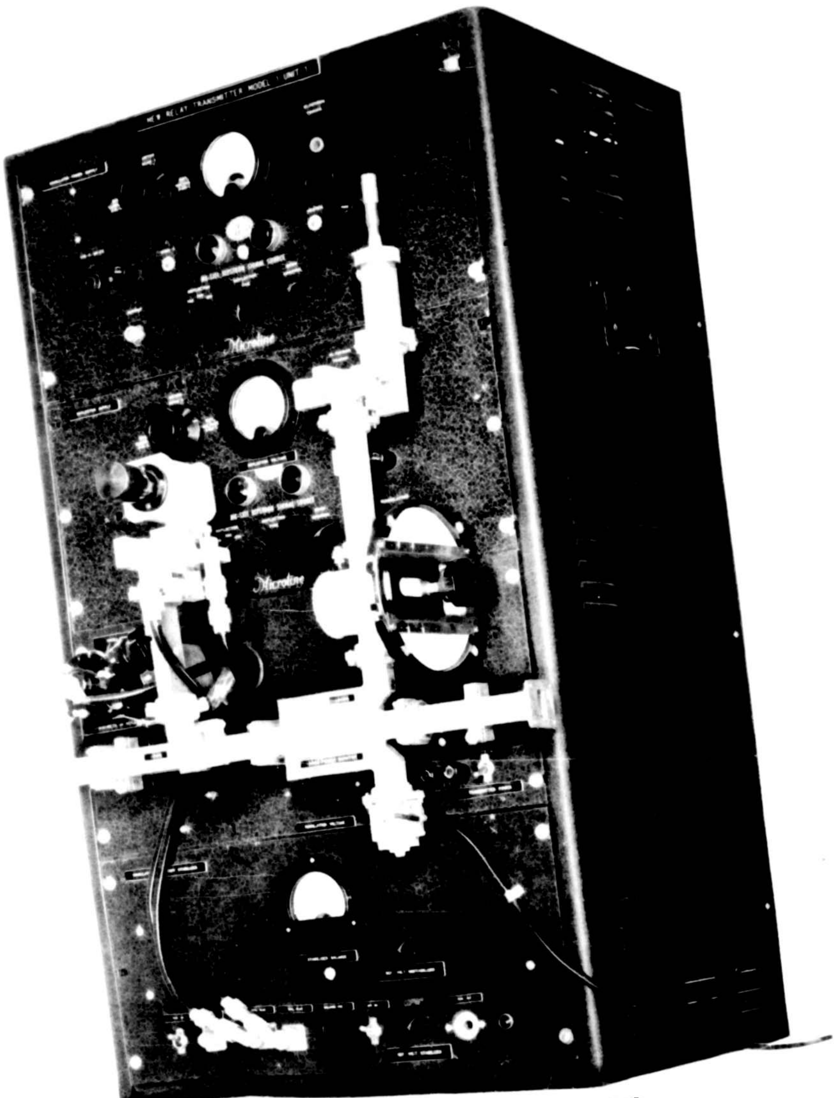


FIG. 1-1 TRANSMITTER LEFT FRONT OBLIQUE

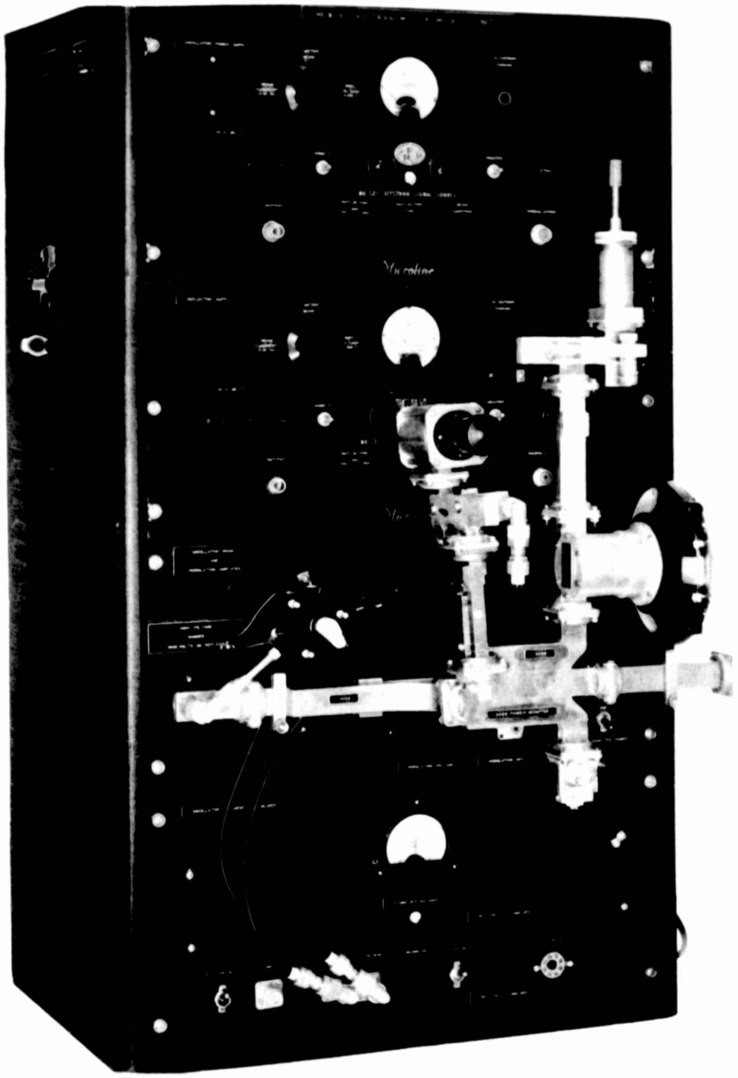


FIG. I-2 TRANSMITTER RIGHT FRONT OBLIQUE

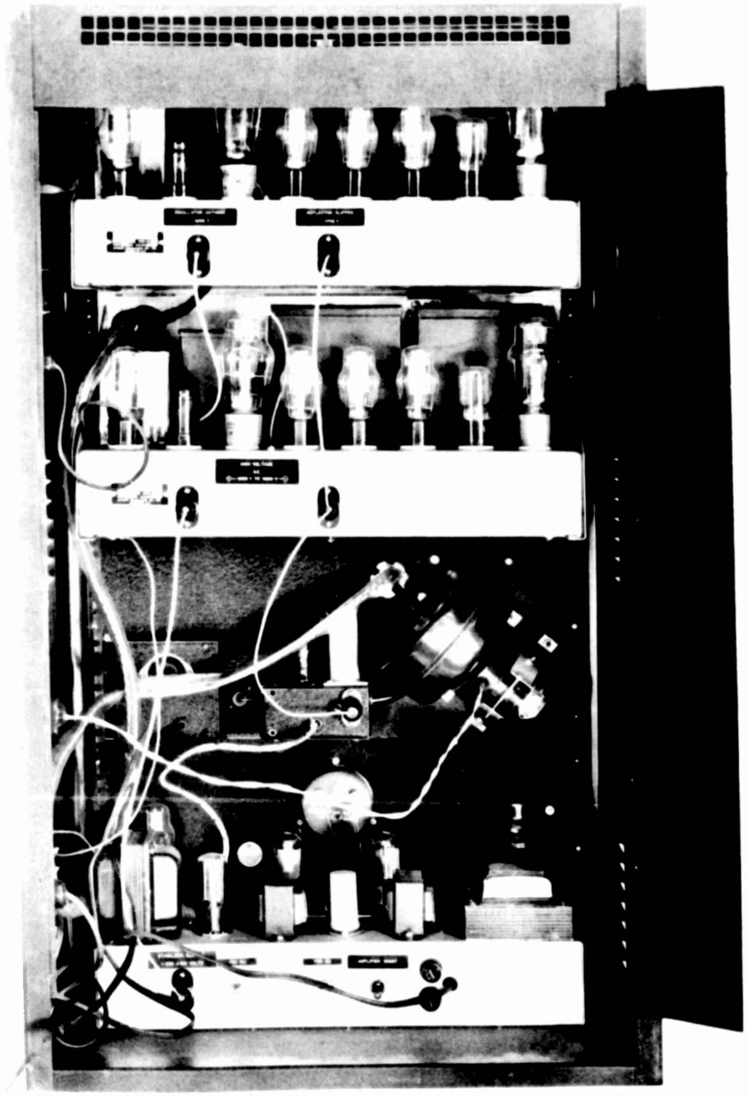


FIG. I-3 TRANSMITTER REAR VIEW

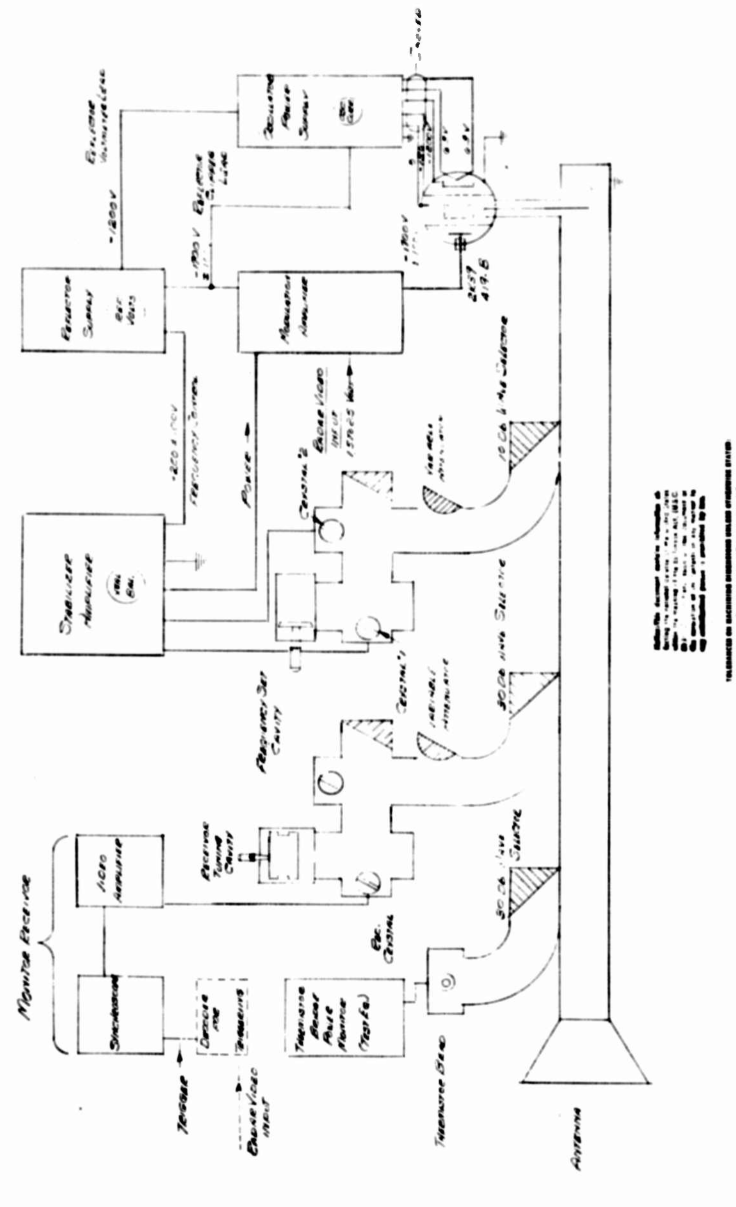


FIGURE I-4 M.E.W. RELAY TRANSMITTER MODEL I

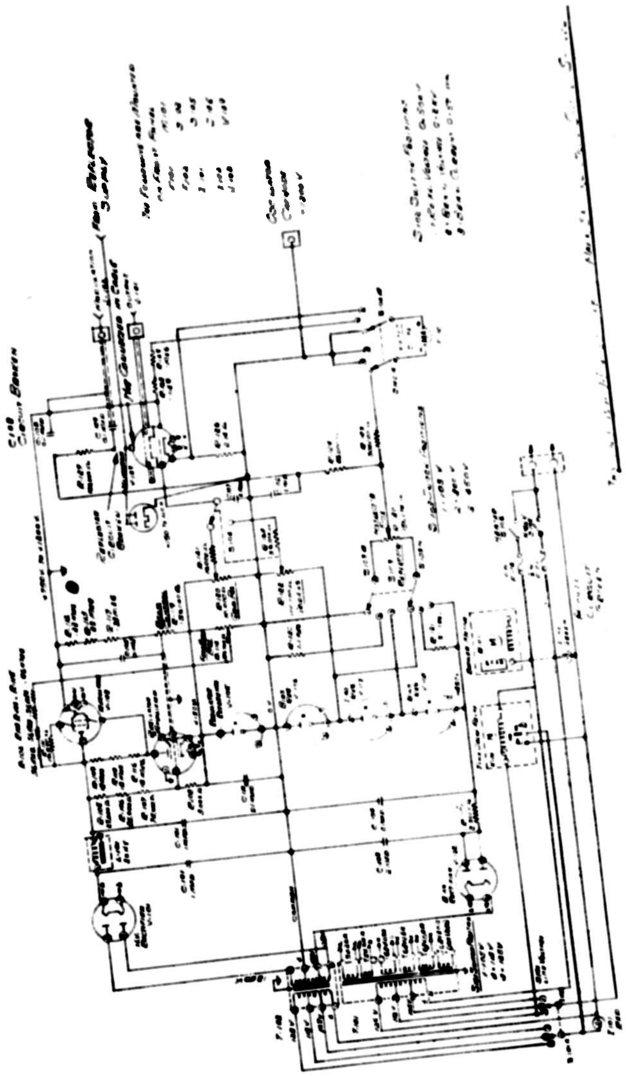


FIGURE I-5 OSCILLATOR POWER SUPPLY FOR M.E.W. RELAY TRANSMITTER

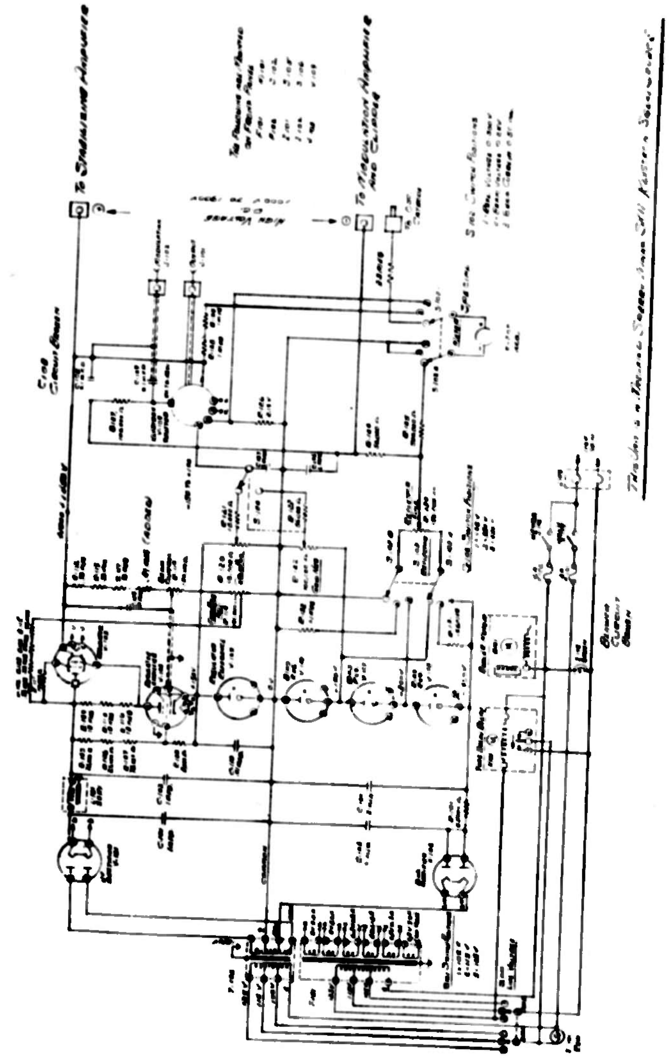


FIGURE I-6 REFLECTOR SUPPLY FOR M.E.W. RELAY TRANSMITTER

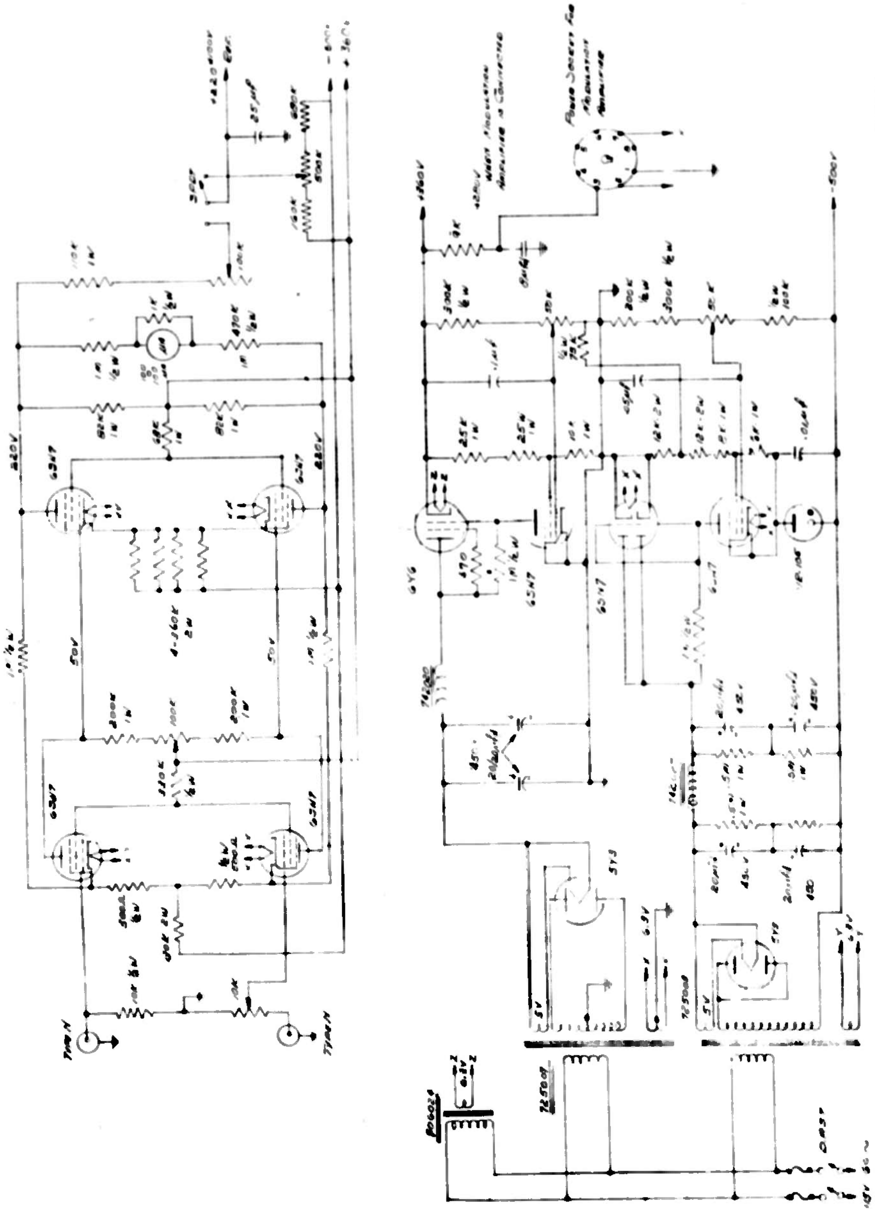
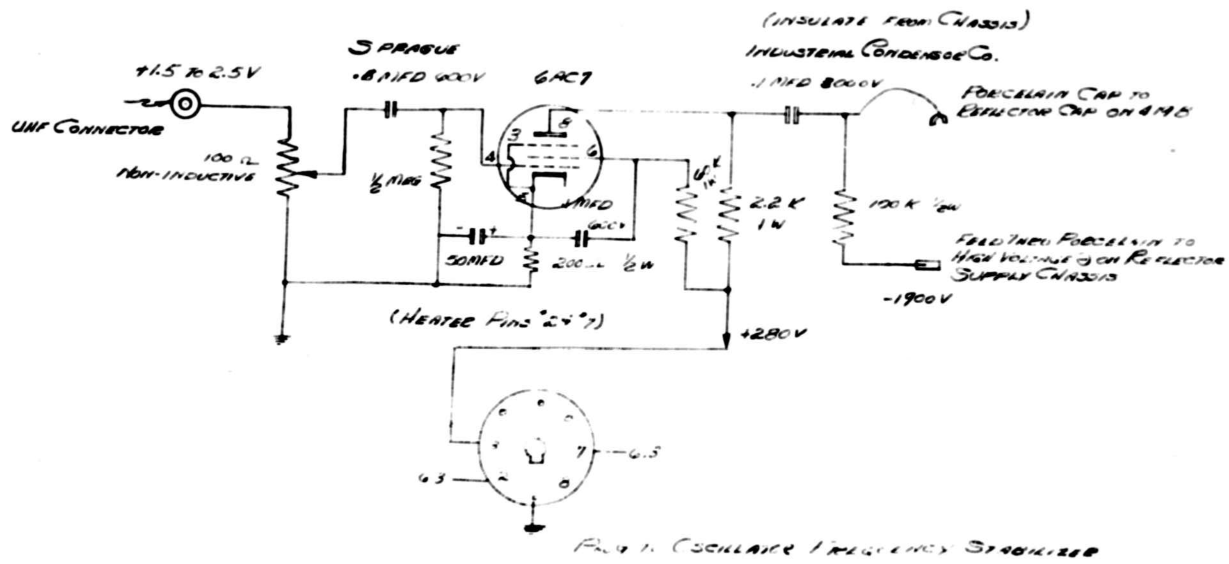


FIGURE I-7 OSCILLATOR FREQUENCY STABILIZER FOR M.E.W. RELAY TRANSMITTER



TOLERANCES ON MACHINING DIMENSIONS UNLESS OTHERWISE STATED:
FRACTIONAL ± 1/64", DECIMAL ± .005", ANGULAR ± 1/2°

FIGURE I-8 SINGLE STAGE MODULATION AMPLIFIER FOR M.E.W. RELAY TRANSMITTER

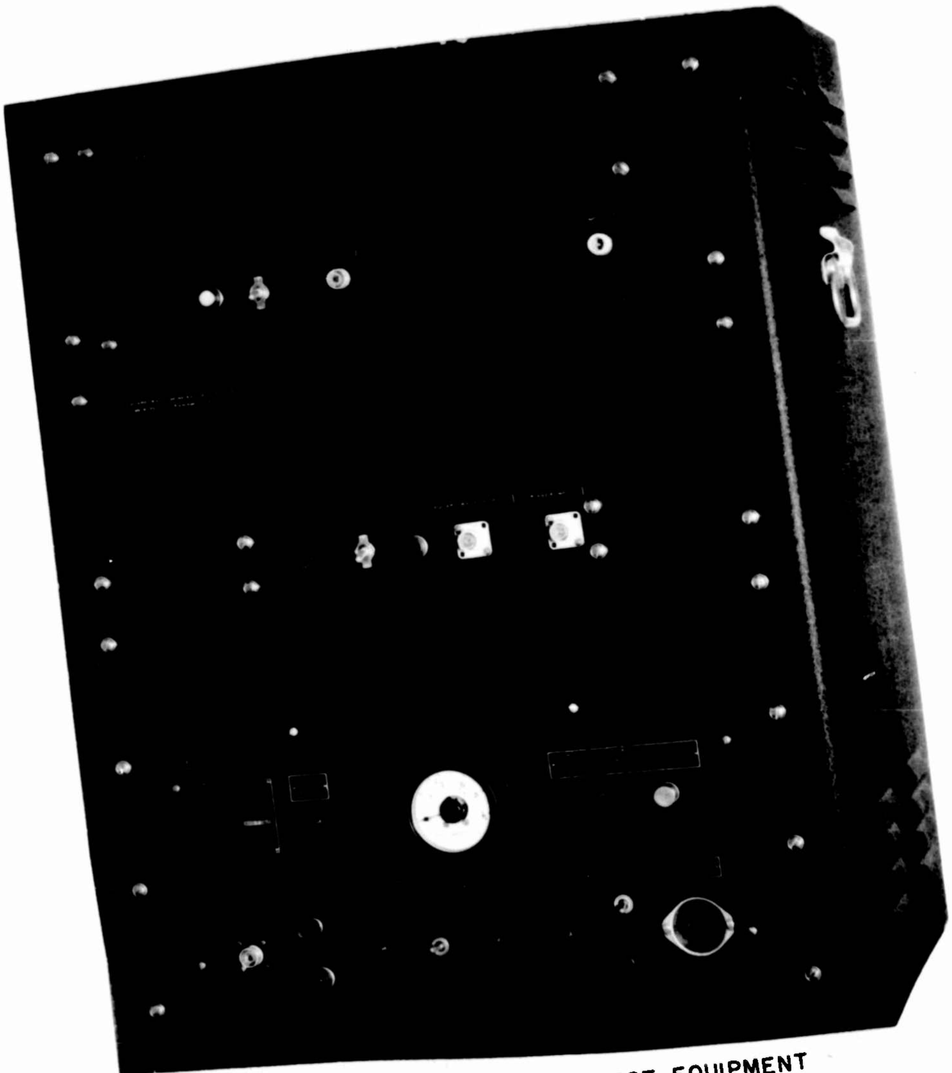


FIGURE I-9 AUXILIARY TEST EQUIPMENT

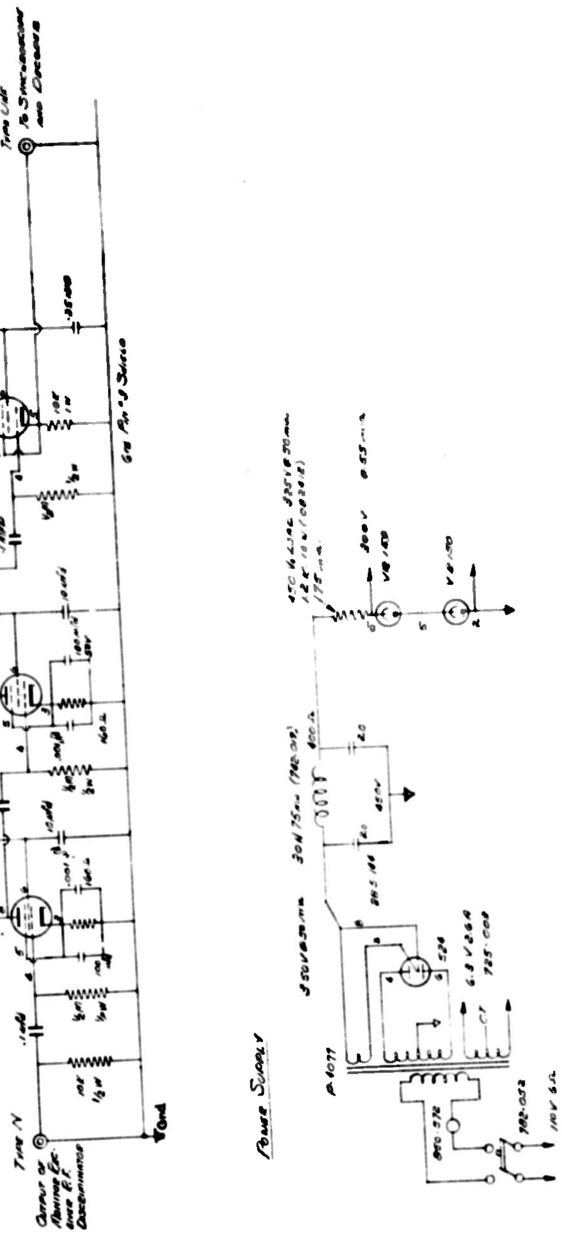
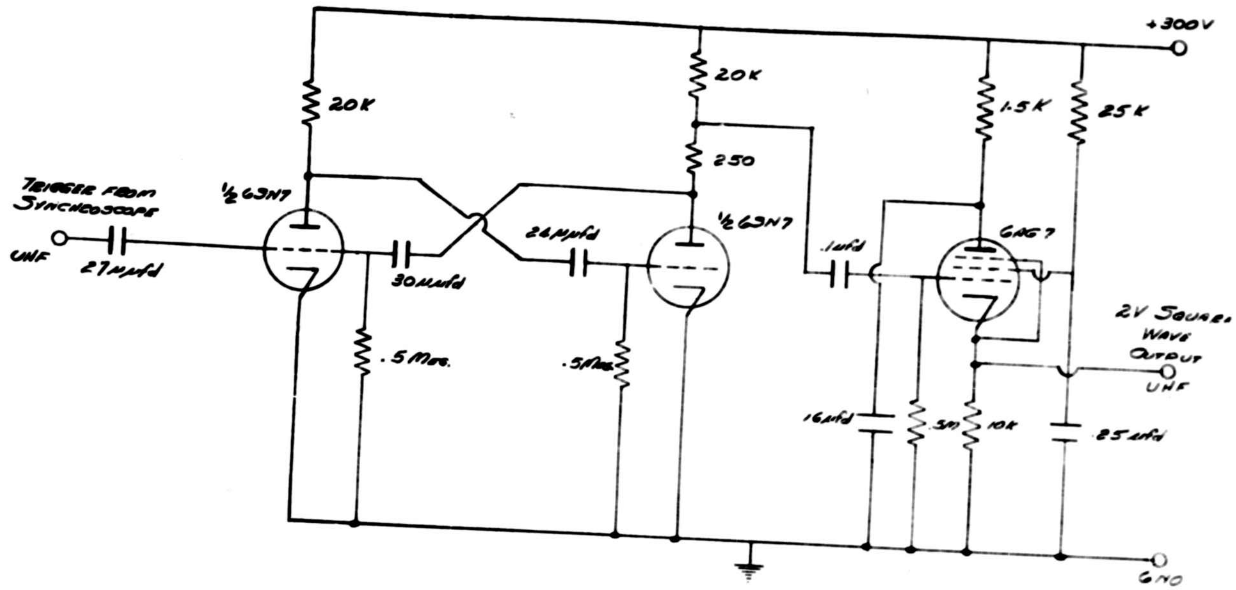
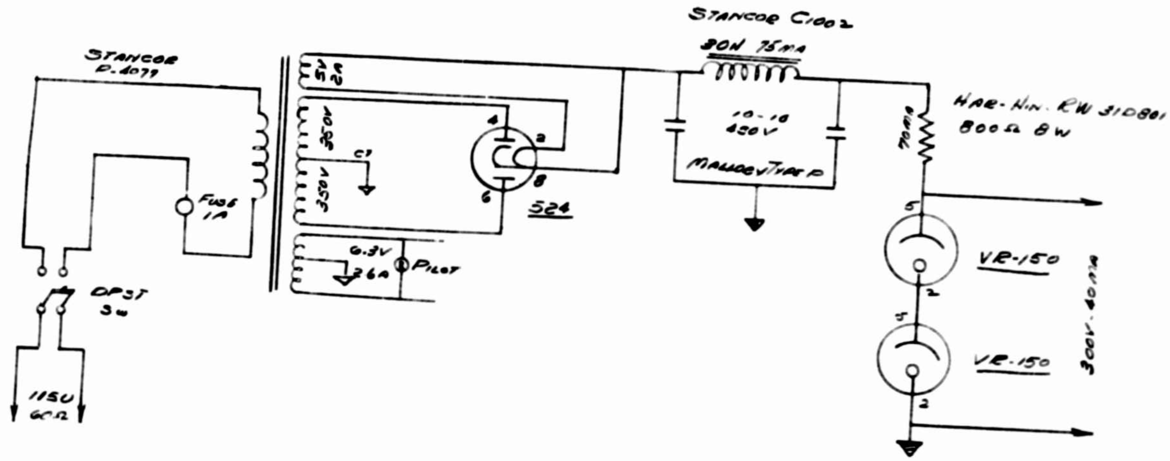


FIGURE I-10 VIDEO AMPLIFIER FOR M.E.W. RELAY MONITOR RECEIVER



TOLERANCES ON MACHINING DIMENSIONS UNLESS OTHERWISE STATED:
FRACTIONAL ± 1/64", DECIMAL ± .005", ANGULAR ± 1/2°

FIGURE I-11 TRIGGERED SQUARE WAVE GENERATOR FOR M.E.W. RELAY TRANSMITTER



TOLERANCES ON MACHINING DIMENSIONS UNLESS OTHERWISE STATED:
FRACTIONAL ± 1/64" DECIMAL ± .005". ANGULAR ± 1/2°

FIGURE 1-12 POWER SUPPLY FOR TRIGGERED SQUARE WAVE GENERATOR

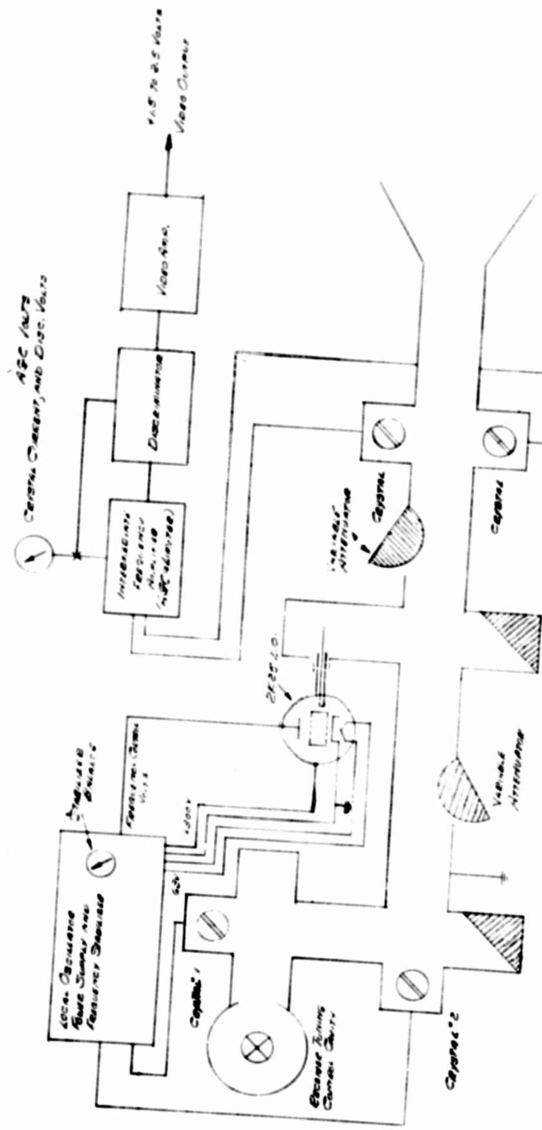


FIGURE I-13 MEW RELAY RECEIVER

FIGURE I-14 RECEIVER TOP FRONT OBLIQUE

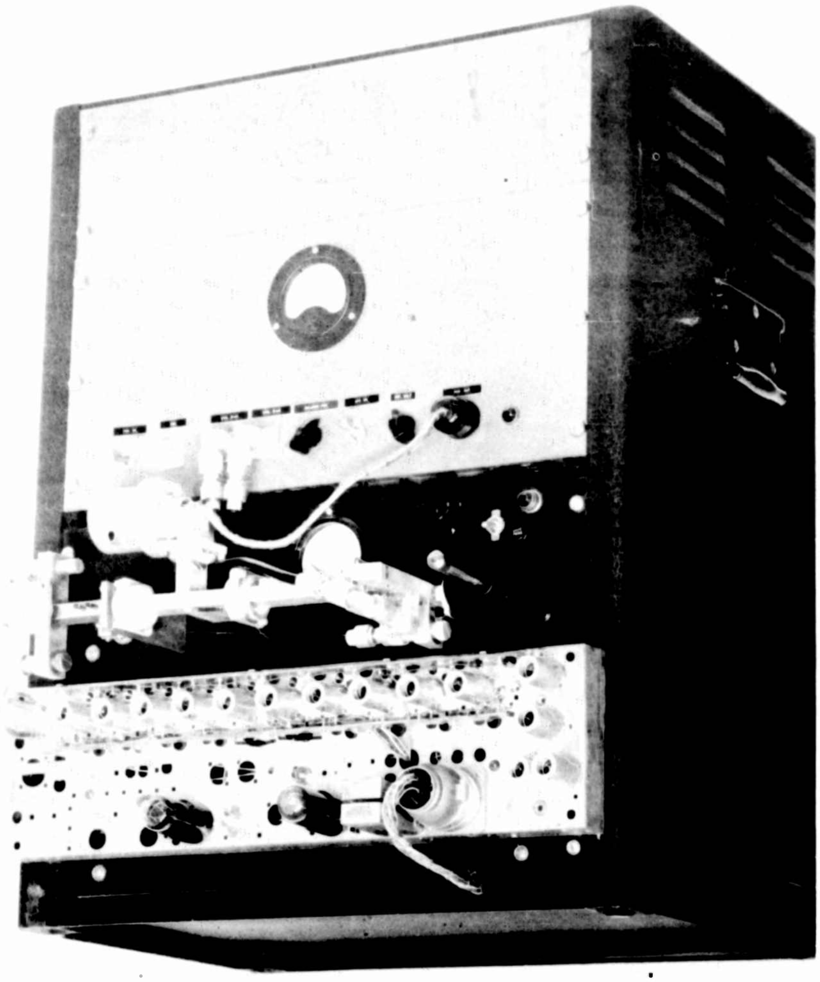


FIGURE I-15 RECEIVER - BOTTOM FRONT OBLIQUE

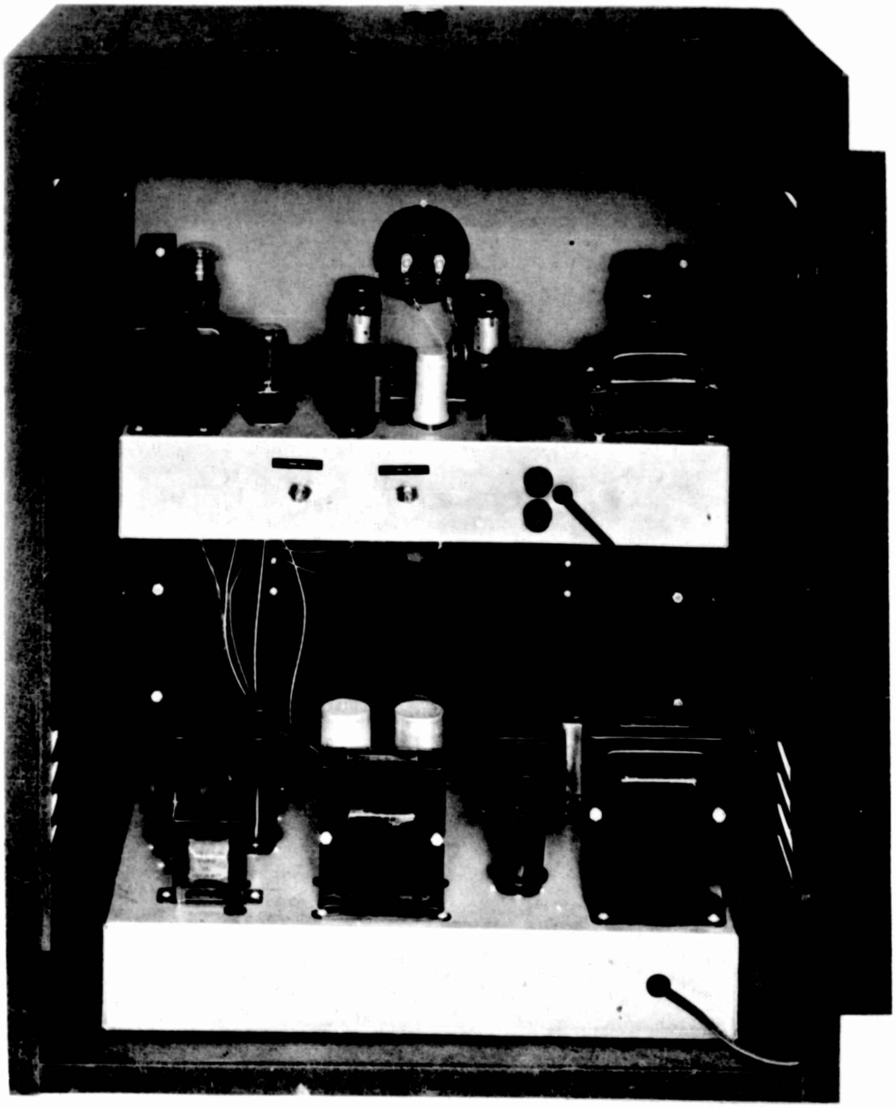
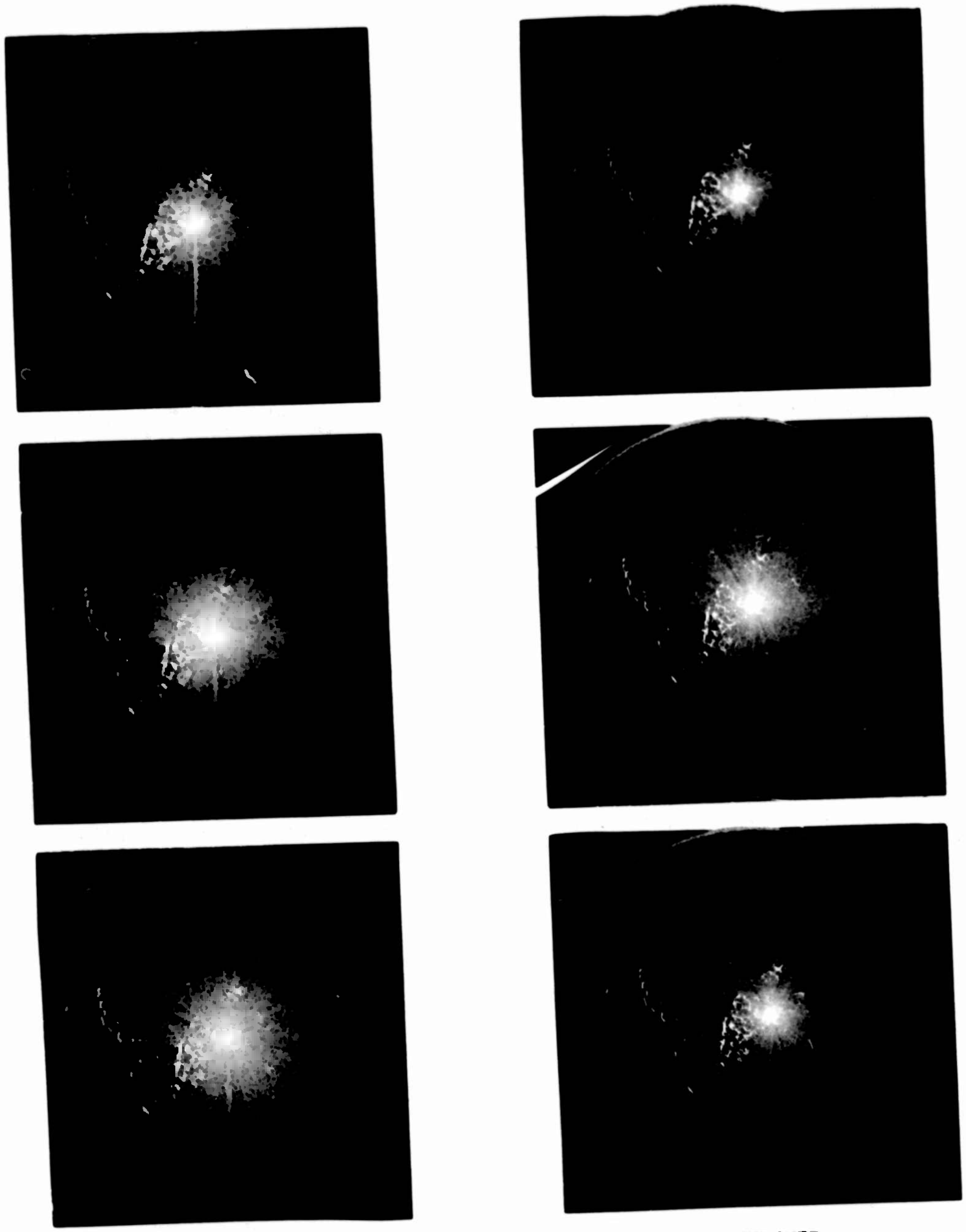


FIG. I-16 REAR VIEW OF RECEIVER



ORIGINAL RELAYED

FIG. I-17 TEST RESULTS

Section 2.

X-BAND FM RELAY RECEIVER

2.1 INTRODUCTION

This section will consider only that portion of the receiver from the input circuit of the i-f amplifier to the video output. The r-f components, AFC, LO, and mixer are covered in Section 1. The block diagram of the complete receiver is shown in Fig. 2.1, and a schematic diagram of the portion to be covered here is shown at the end of the section in Fig. 2.2. Photographs of the complete receiver are shown in Fig. 1.14 and Fig. 1.15 in Section 1. A few general remarks regarding the design of FM receivers and limiters will be made first, and this will be followed by a more detailed description of the X-band relay radar receiver.

Only two receivers have been built for the ground relay system. Both these receivers are similar electrically but differ slightly mechanically. They are to be considered only as laboratory models, however, which were hastily assembled to get some idea of their electrical performance, as well as the overall system performance.

The i-f amplifier chassis and general layout is similar to the Mk 56 receiver, since most of the parts, coil forms, hardware, and chassis were readily available. The power supply was made electronically regulated only because it was expected to use these receivers in truck installations for field strength measurements where extreme gain stability is essential. Whether regulated supplies should be used in the final design or not has never been given any thought. This will depend on the choice of the power source in the field as well as expected load variations. No thought has been given the mechanical layout of the receiver as the final packaging of the system was not known.

2.2 GENERAL REMARKS REGARDING THE DESIGN OF FM RECEIVERS

2.21 Noise Considerations

A good insight to FM receiver performance can be gained by reference to the work of M. G. Crosby.* An extension of his work, by H. M. James**, includes c-w and pulsed interference problems. In general, they have shown that at low signal levels or strong interference the AM system is to be preferred to the FM system. Frequency modulation, however, does show better performance at high signal levels, and, in particular, a large deviation ratio FM system has a decided advantage over AM against strong impulse noise.

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* M.G.Crosby, Frequency Modulation Noise Characteristics, Proc. I.R.E., Vol.25, pp 472-514, April, 1937.

** H.M.James, Notes on Noise, Interference, and Distortion in FM Transmissions, RL 43-3/23/45.

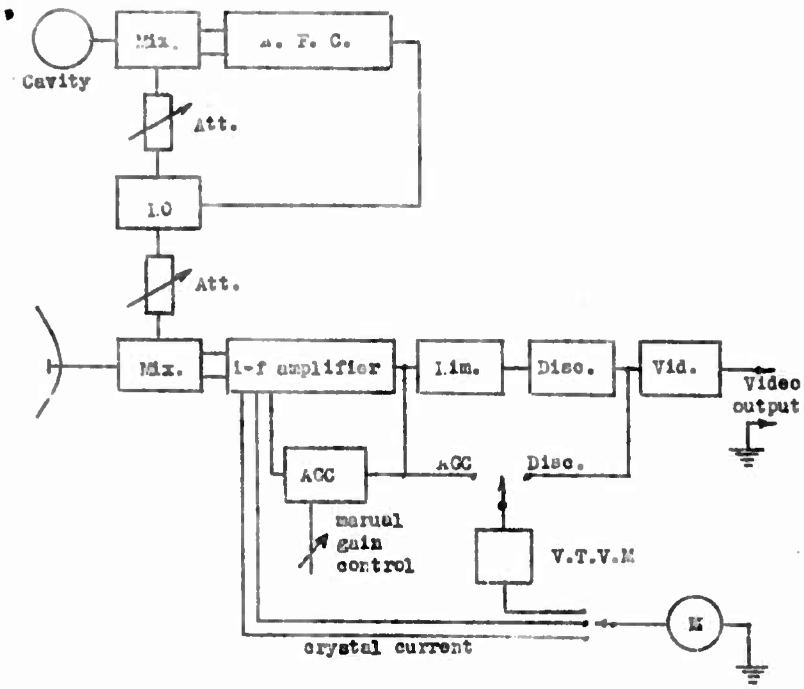
Mix.
A. F. C.
Cavity
Att.
L.O
Att.
Mix.
i-f amplifier
Lim.
Disc.
Vid.
Video
output
AGC
AGC
Disc.
manual
gain
control
V.T.V.M
M
crystal current

Fig. 2.1

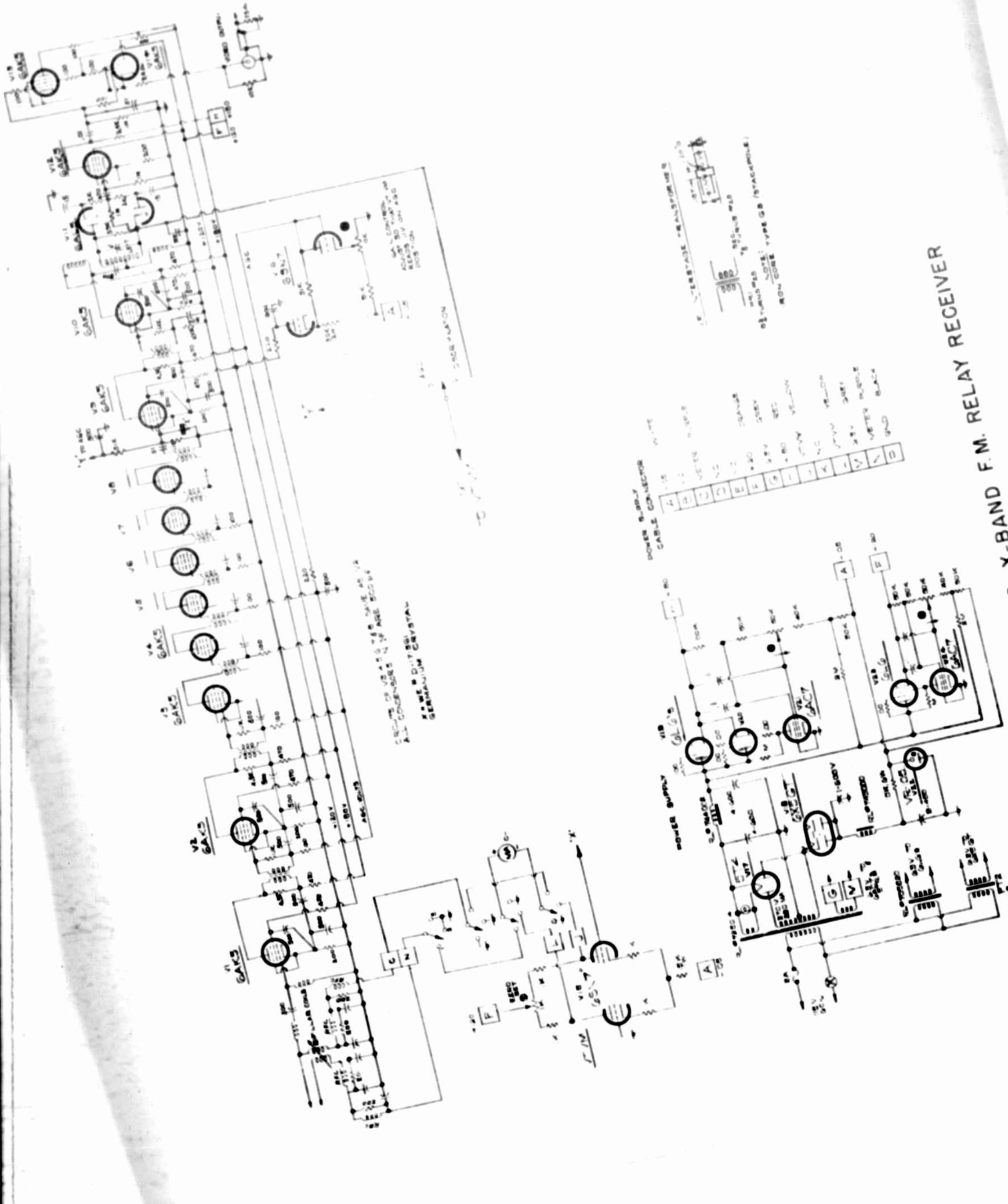

FIGURE 2-2 X-BAND F.M. RELAY RECEIVER

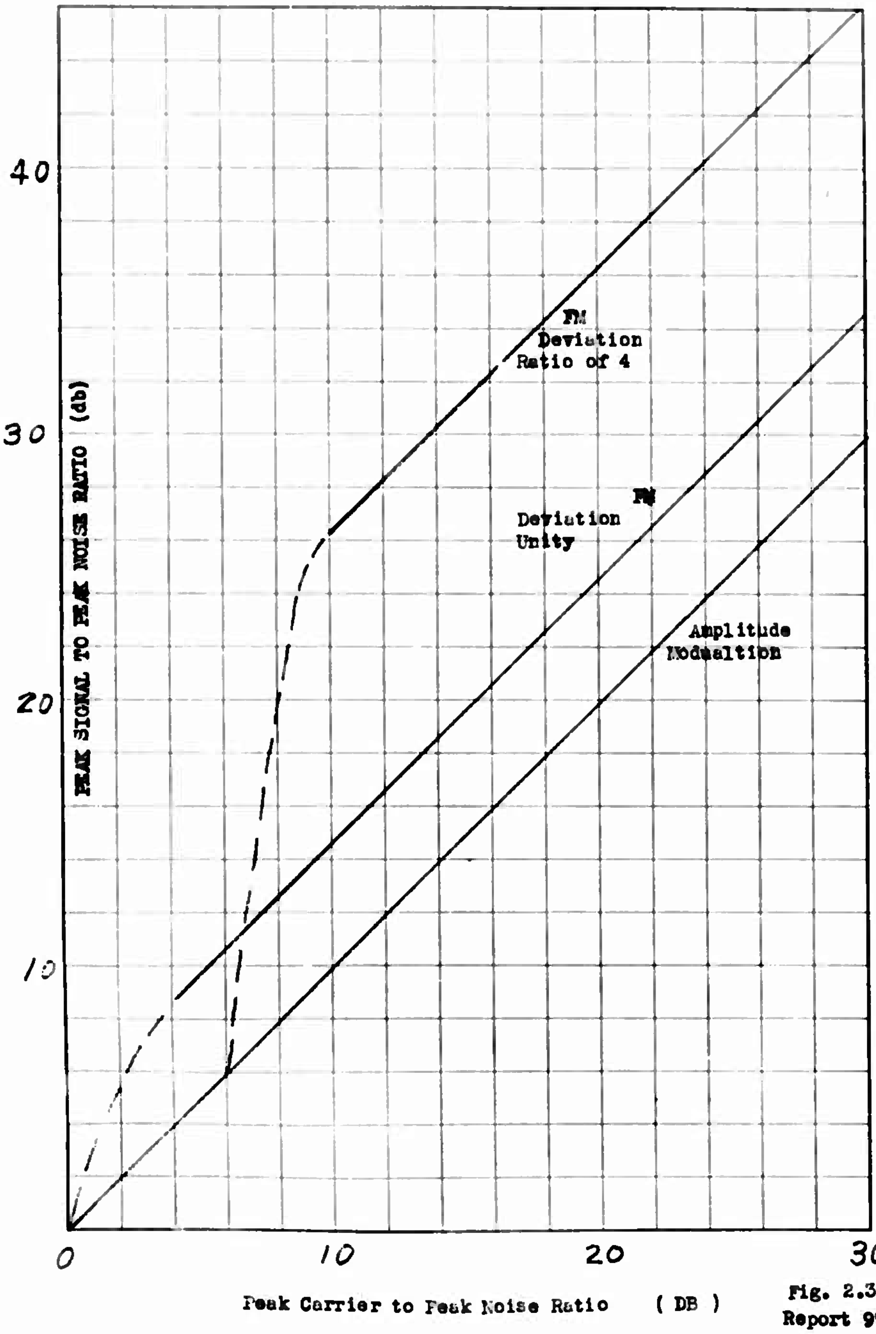

Fig. 2.3
Report 977,Sec.2

The relative performance of FM and AM receivers on the basis of peak signal and peak fluctuation noise is shown in Fig. 2.3. These curves were derived by passing the noise and FM carrier through a limiter and discriminator mathematically, but have been verified experimentally. Here the peak carrier to peak noise ratio has been plotted as abscissa and the peak signal to peak noise ratio plotted as ordinates. The term "carrier to noise" ratio is used to denote the signal to noise ratio before the second detector in an AM system, and before the limiter and discriminator in an FM system, whereas "signal to noise" refers to the signal to noise ratio after detection. In an AM receiver, the "signal to noise" ratio is the same as the "carrier to noise" ratio, at least when the latter is greater than zero, as is indicated in Fig. 2.3 by the straight line through the origin at 45°. However, in FM, the "signal to noise" ratio is greater than the "carrier to noise" ratio at large signal levels. In Fig. 2.3 we see that at a carrier to noise ratio on the order of 30 db, an FM receiver with a deviation ratio of 4 has a signal to noise ratio approximately 16 db above the signal to noise ratio of the AM receiver, and that with a deviation ratio of unity the improvement is about 5 db. This improvement is seen to be constant for high signal levels, but then drops suddenly to no improvement as the signal level is decreased. The point at which the sudden drop crosses the AM curve is called the "improvement threshold". Note that with a large deviation ratio there is a greater improvement in signal to noise but that the required signal level to reach the "improvement threshold" is less for the small deviation ratio.

This is really an over-simplified picture of the relative performance of FM and AM receivers as it considers only peak signal to noise conditions. The relative performance will be slightly different for rms noise and impulse noise conditions, and other factors, such as interference and distortion, come into the picture to alter the appraisal of the overall system from that given here. The reader is urged to refer to the work of Crosby and James mentioned earlier for a more complete discussion of these relative merits.

2.22 Gain Requirements

We shall explicitly assume that the FM receiver is to be used in a system where the carrier to noise ratio is always well above the improvement threshold, as this is the only case where the better performance of FM over AM is realized. However, the limiter and the gain of the i-f amplifier should be so designed that the ratio of the limit level to peak noise equals the improvement threshold. Let us denote the ratio of peak carrier to peak noise by p, and the theoretical improvement threshold ratio by t. If $p > t$, the actual improvement threshold will no longer be t, but will approach p, as it is necessary to limit in an FM system in order to realize any improvement at all. Further, there is no point in providing so much gain that $p < t$, as t, for a given deviation ratio, represents a lower limit that cannot be bettered, in fact, that is the definition of t. Thus the gain of the i-f amplifier should be enough to bring the tube and Johnson noise of the first stage to a level such that the ratio of the limit level to peak noise is equal to the improvement threshold.

The effective rms noise at the grid of the first tube is arrived at just as in AM receiver design and is given by

$$E = \sqrt{4KTRB} \qquad (2.1)$$

where K = Boltzmann's constant = 1.380×10^{-23} joule per °K
T = absolute temperature °K
R = effective resistance in grid circuit
B = overall 3 db bandwidth*

If the input circuit is a transitionally coupled double tuned circuit loaded on the primary only, then

$$R = 2R_1 \frac{C_1}{C_2} \qquad (2.2)$$

where R_1 is the primary loading, C_1 the primary capacitance, and C_2 the secondary capacitance. This is the most usual case in input circuit design when the input tube is a pentode, although sometimes single tuned circuits are also used at 60 Mc/sec. The value of noise given by equation (2.1) is then multiplied by the square root of the noise figure of the i-f amplifier. This noise figure must be estimated in the early stages of design but can be measured** after the first model has been built.

To measure the ratio of limit level to peak rms noise, the following procedure suggested by Y. Beers can be used. A block diagram of the required setup is given in Fig. 2.4. The signal generator should have calibrated output and means should be provided for reducing the gain of the receiver a known amount on the order of 30 db. The power measuring device should respond to rms noise power, and sufficient accuracy for this measurement can be obtained by the use of a linear detector.***

With SW2 at "a", the receiver gain is adjusted so that the thermal noise will develop, say, 0.1 volt at the grid of the first limiter tube. SW2 is now thrown to "b" and the reading of the diode current meter noted. If SW1 is now closed and the signal generator output adjusted so that the reading of the diode current meter is doubled, the output of the signal generator, P_1, will be equal to the rms noise power of the receiver. The gain of the receiver is then reduced by a known amount, α, on the order of 20 or 30 db. With SW2 on "a", the output of the signal generator is increased to the value P_2 until the signal just limits, as indicated by the V.T.V.M. on the discriminator output. This, of course, requires that the signal generator be detuned from the crossover frequency of the discriminator. Since the gain of the receiver has been reduced by α, the peak signal power required to just limit will be given by $2\alpha P_2$. Thus we have, in db,

---

* This is an approximation as the noise bandwidth and the 3 db bandwidth are not the same; actually there is very little difference in a multistage amplifier. The variation of R_1, in Eq. (2.2), with crystal current is a much more important factor.

** See Y. Beers, Receiver Noise Figures and Their Measurements, RL #746.

*** See Fig. 15 in RL 746.

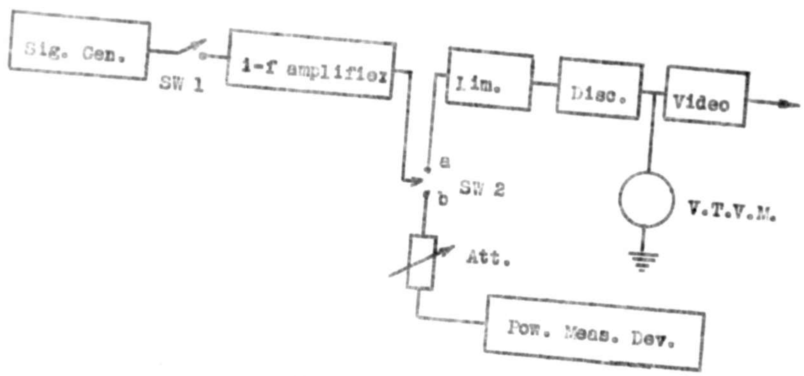

Fig. 2.4

$$\frac{C\ (\text{peak})}{N\ (\text{rms})} = 10 \log \frac{2\alpha P_2}{P_1} \qquad (2.3)$$

which is the ratio of the limit level to the rms noise power at the limiter. The ratio of the limit level to the peak noise power can be obtained by multiplying P_1 by the "crest factor", which is defined as the ratio of peak noise to rms noise. The concept of a definite value of "peak" noise cannot be justified on a theoretical basis because of the positive probability that at some time arbitrarily large noise peaks will occur. However, it has been found that, for practical purposes, the ratio of peak noise to rms noise is about four.* Therefore, we have

$$\frac{C\ (\text{peak})}{N\ (\text{peak})} \approx 10 \log \frac{2\alpha P_2}{4^2 P_1} \approx 10 \log \frac{\alpha P_2}{8P_1} \qquad (2.4)$$

By adjusting the i-f amplifier gain of the limit level, the above ratio can be made equal to the "improvement threshold". When this has been done, the rms noise voltage at the grid of the limiter can be determined by,

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* See K. G. Jansky, An Experimental Investigation of the Characteristics of Certain Types of Noise, Proc. I.R.E., Vol 27, pp 763-768, Dec. 1939. For a discussion of "crest factor", see V. D. Landon, The Distribution of Amplitude with Time in Fluctuation Noise, Proc. I.R.E. Vol. 29, pp 50-55, Feb. 1941

$$\text{rms noise voltage at limiter} = \frac{E}{\alpha}\sqrt{\frac{P_1}{P_2}} \qquad (2.5)$$

where E is the rms voltage required to just limit. The value of E can be found by measuring the output of a signal generator applied to the grid of the limiter. The gain of the i-f amplifier can thus be decided upon, knowing the rms noise at the input of the receiver and the noise required at the limiter.

2.23 Bandwidth Requirements

A rigorous analysis of the frequency spectrum of a pulsed frequency modulated signal, such as that shown in Fig. 2.5, has been made by H. M. James.* Only a brief discussion of bandwidth requirements will be given here. James' analysis shows that for $\delta \ll \tau$, where δ is the pulse length and τ the period between pulses, the spectral distribution can be approximated closely by considering the complete wave train consisting of three parts:

(a) a continuous wave of frequency f_c,

(b) a sequence of pulses of frequency f_c, duration δ, with intervals of $(\tau - \delta)$, having equal amplitude and opposite phase to the above frequency, and

(c) a sequency of pulses of frequency f_p, duration δ, with interval $(\tau - \delta)$, filling in the gaps in the wave train created by the interference of (a) and (b).

The spectrum will correspondingly consist of:

(a) a few strong lines of frequency f_c, carrying most of the energy,

(b) a distribution of lines around f_c, with intensity varying like that for a pulse of duration δ, carrying a fraction δ/τ of the total energy, and

(c) a distribution of lines near f_p, with an intensity distribution like that for a pulse of duration δ, carrying a fraction δ/τ of the total energy.

The last two distributions will overlap and interfere as already noted. Thus the bandwidth must be equal to the frequency deviation plus that required to pass a pulse of duration δ.

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* H. M. James, Harmonic Analysis of Rectangular FM Pulses, RL Report #43-5/25/45

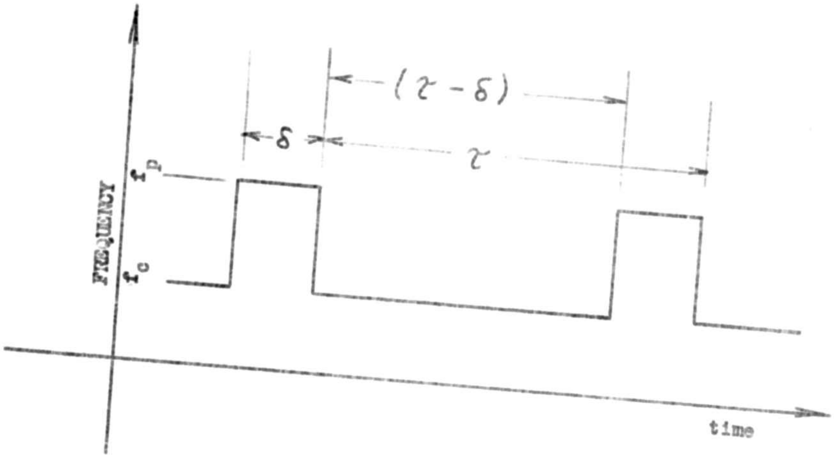


Fig. 2.5

2.24 Limiter Design

The function of the limiter is to remove amplitude modulation from the carrier in an FM receiver. The limiter can be either one or two stages, the two-stage, or "cascade limiter", being preferred. The problem to be met in its design is first to have the output constant over a large range of input voltages, thus reducing amplitude modulation; and second, to have the time of operation short enough to effectively reduce impulse noise.

2.241 Single Stage Limiters

The usual design of an FM limiter is based on "grid limiting" and takes the form of Figure 2.6. If the applied voltage is large enough to draw grid current, the voltage developed across the resistor R will charge condenser C negatively. During the negative half of the cycle, the condenser charge leaks through R. If $RC \gg 1/f_c$, the negative voltage thus developed across R at the grid of the limiter will be proportional to the applied input voltage. Under those conditions, the range of the effective voltage change in the grid circuit, responsible for plate current change in the plate circuit, will be between the tube cutoff and that value at which the tube draws grid current, i.e., grid-base. Thus the effective grid swing will be constant and independent of the magnitude of the applied voltage as long as the applied voltage is large enough to draw grid current.

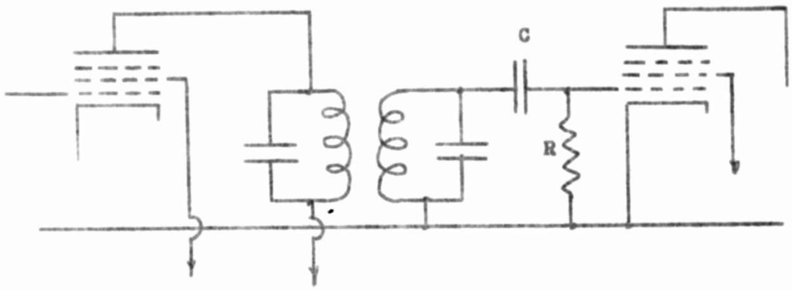


Fig. 2.6

By lowering the plate and screen voltage on the limiter stage, the range of the effective grid voltage swing can be reduced. This not only reduces the voltage required for cutoff, but can also cause plate current saturation, or "plate limiting", if the plate load resistor is large enough for its loadline to fall well below the knee of the tube characteristics. In wide band limiter design, the plate load resistor cannot be made large enough to take advantage of plate limiting.

Even though the effective grid swing is constant, the output of the limiter is not necessarily constant. This is because with large input voltages enough bias can be developed to operate the limiter class "C". As the plate current angle of flow is reduced with increase of input voltage, the harmonic content of the plate current will increase, and the fundamental component, which will be the only one passed by the tuned discriminator, is therefore reduced. This results in a limiter that will limit over a certain range of input voltages but will have reduced output at higher input voltage levels. The above discussion assumes that the effective grid swing is constant. Actually, this is not the case. The effective grid swing will be constant only if R is much greater than the grid to cathode resistance when the grid is conducting. Thus, by reducing the value of R, the grid swing will increase slightly with increased signal level, and a critical value of R can be found where this increased grid swing just balances the loss in output due to the smaller angle of flow of plate current. This results in a limiter that can be made to have relatively constant output over a wide range of input levels. This is still not a completely satisfactory limiter, however, because of impulse noise considerations.

2.242 Impulse Noise and Time Constants

The value of C should be such that the time constant, RC, is very much greater than $1/f_0$ (f_0 being the intermediate frequency) in order to assure reasonable efficiency in developing bias voltage, but not so large that the circuit cannot recover shortly after a noise impulse. The shortest impulse, or the fastest amplitude modulation, that can be passed will be determined by the bandwidth of the i-f amplifier. When a step function is passed through an amplifier, the rise time, from 10% to 90% of its amplitude, is approximately .7/B seconds. The approximate duration of the rise and fall of the shortest impulse that can be passed by the amplifier is, therefore, approximately 2/B seconds. If we now ask that the time constant be, say, 1/6 the duration of the shortest impulse, we have

$$\frac{1}{3B} > RC > \frac{1}{f_o} \qquad (2.6)$$

In particular (2.6) requires $3B < f_o$; often this condition cannot be met in wide band receivers where it is not uncommon that $3B = f_o$. If the bandwidth is fixed by the required deviation or transient response, the only thing left to do is to increase the value of the intermediate frequency.

Usually wide band receivers have deviation ratios greater than unity, and, therefore, the video will be too narrow to pass as fast on impulse as the i-f amplifier. If the limiter time constant is chosen on the basis of video rather than i-f bandwidth, a large noise impulse passed by the i-f amplifier may overload the first video stage. Thus it is desirable to follow the above expression even in the design of receivers for large deviation ratios.

Another complication in the design of grid limiters is the change of loading on the coupling circuit caused by the grid when conducting. This calls for a large time constant so that the angle of grid current flow is small. In the interests of fast action, a large time constant cannot be used. The effect of this additional loading can be reduced, however, by using a small L/C ratio in the tuning circuit, resulting in a decreased load resistance for a given bandwidth. This large C will also swamp out any changes in the input capacitance of the tube. The above difficulty is not as likely in broad band receivers where the Q of the circuit is less than ten.

Even though single stage limiters have found considerable use, the compromise made between fast action, constant output, and loading and gain of the previous stage, indicates the desirability of two stage limiters, or "cascade limiters".

2.243 <u>Cascade Limiters</u>

The "cascade limiter" consists essentially of two single stage limiters in series. Here the first stage is designed for most effective impulse noise elimination, with no regard for constant output, and the second stage is designed for constant output, leaving the burden of removing impulse noise to the first stage. The time constant of the second stage is not important and can be made large. This will result in most constant output since the first stage does limit to a certain extent, and there will be little danger of the output's falling off at larger input levels. To assure that the output of the last stage will not fall off, the gain of the first stage should be kept low.

It is necessary that the short time constant be in the first limiter so that fast variations of the amplitude will be removed before being applied to the slow limiter. If this were not done, a noise impulse could develop enough bias to reduce the signal level below the point where the second limiter could follow.

2.3 DESCRIPTION OF THE RECEIVER

2.31 Input Circuit

The input circuit is a π equivalent of a double tuned circuit and was used because of its ease in alignment and of obtaining separate crystal current outputs for each crystal. This circuit, Fig. 2.7, can be arrived at by the sequency of equivalent circuits shown in Fig. 2.8. Although the crystals are shown "opposed", the i-f output voltages will be in phase, since the r-f voltage applied to them is out of phase. The bandwidth of the input circuit was designed to be 20 Mc/sec when transitionally coupled with R_1 equal to 200 ohms.* The i-f impedance of crystals is usually considered to be on the order of 300 ohms, which, with two crystals in parallel would make R_1 equal to 150 ohms. However, if possible, it is considered good practice to design the input circuit to be transitionally coupled at a higher resistance. This is done to minimize the change of bandwidth** of the input circuit caused by the change of crystal impedance with crystal current.

In Fig. 2.8, R_1 and C_1 are the effective resistance and capacitance of the primary, and C_2 is the effective capacitance of the secondary. The values of L_a, L_b, and L_c can be obtained from L_1, L_2, and M by the following well-known transformations:

$$L_a = \frac{L_1L_2 - M^2}{L_2 - M}$$

$$L_b = \frac{L_1L_2 - M^2}{L_1 - M} \qquad (2.7)$$

$$L_c = \frac{L_1L_2 - M^2}{M}$$

A helpful procedure for aligning a π circuit follows:

(a) Calculate the resonant frequency of L_b and C_2 of Fig. 2.8b. With a high impedance signal generator set this frequency,

---

* The constants of Fig. 2.8a are best arrived at by the use of the curves given in a report by C. P. Gadsden, RL#61-11/17/44, Flat Flat Coupling for the Double Tuned Circuit.

** See Fig. 3 of H. Wallman's report on Stagger-damped Double-tuned Circuits RL #539.

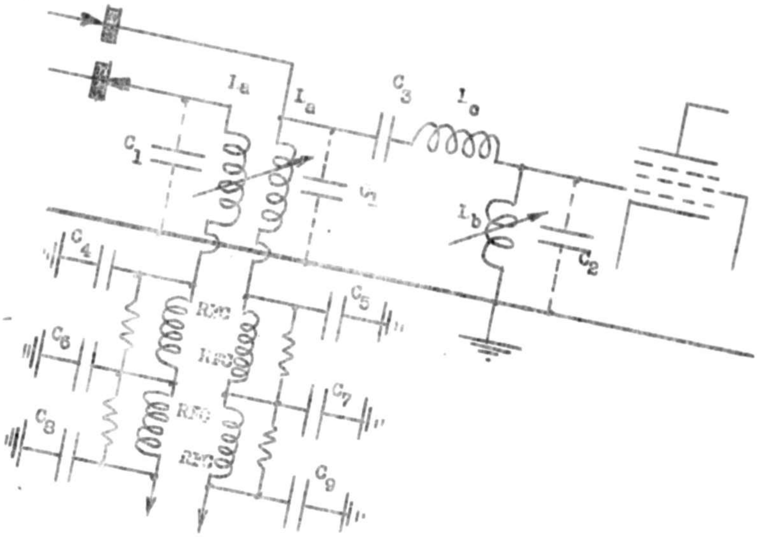
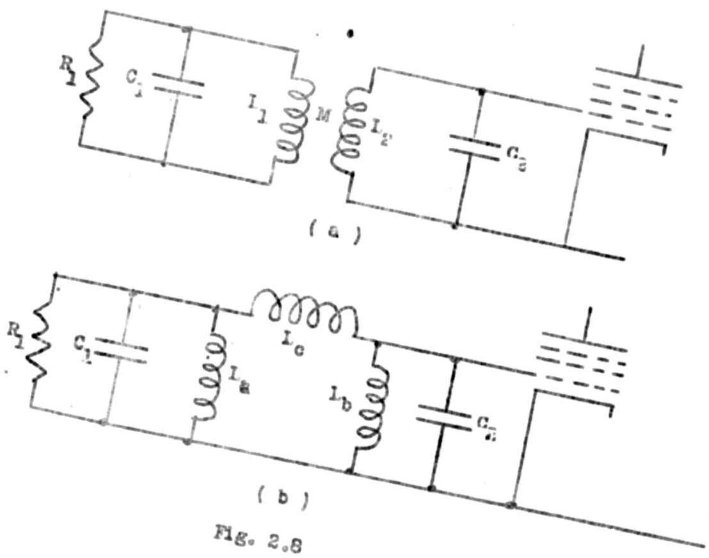


Fig. 2.7



(a)

(b)

Fig. 2.8

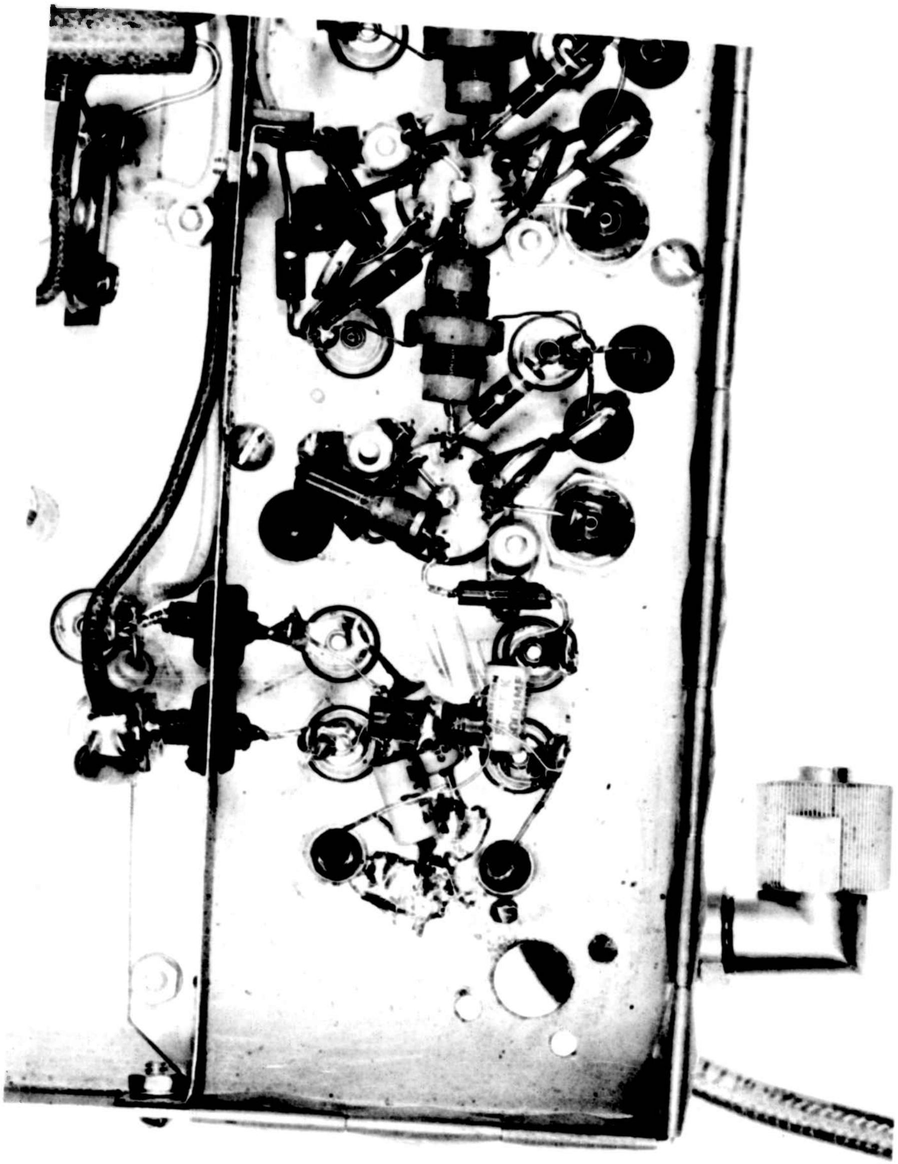


FIGURE 2-9 FM X-BAND RECEIVER - INPUT CIRCUIT

trim L_b to resonance in the chassis to be used with a J.N "center capacitance" tube. See Fig. 2.10a.

(b) Leaving L_b fixed as determined in (a), trim L_c so that the resultant overall circuit is resonant at f_c, where

$$f_c = \frac{f_2}{\sqrt{1-k^2}}, \quad (f_2 \neq f_0)^{*} \quad \text{See Fig. 2.10b.}$$

(c) With the primary circuit properly loaded, trim L_a so that the pass band, as seen with a wobbulator and CRO, is "flat". See Fig. 2.10(c).

Because of the distributed capacitance of L_c, it may be necessary to change the value of L_b slightly when trimming L_a to realize the desired passband. Usually this can be accomplished with no difficulty. The coils can now be taken out of the chassis and measured on the Q-Meter to facilitate duplication in production. It should be entirely possible to make the input circuit "fixed tuned".

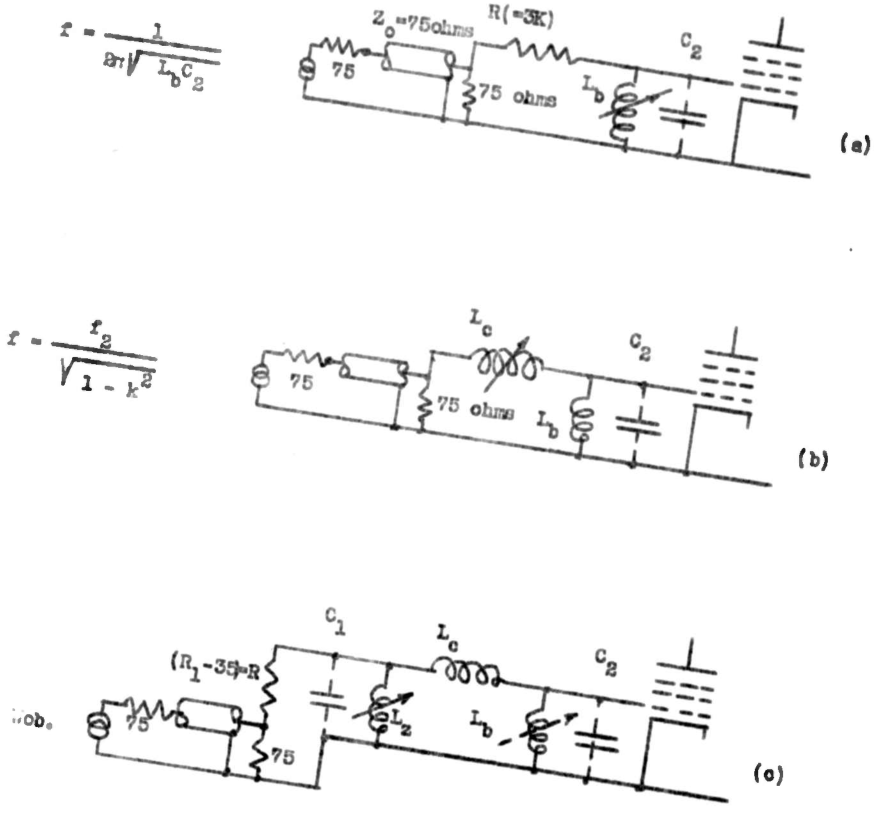
The impedance reflected from the primary to the secondary, as given by equation (2.2), is about 1000 ohms. This results in about 13 μ volts Johnson noise, and assuming an optimistic noise figure of 5 db, (the noise figures of these receivers have never been measured), will give about 24 μ volts noise at the grid of the first tube.

It is urged that the mixer be fixed mechanically to the receiver chassis rather than being connected by coaxial cable as was done here. Although this does not make as flexible a unit, a much more complicated mixer is being used successfully on the Navy Mk 56 Fire Control Radar which is subjected to gun shock. Eliminating the cables is worthwhile in itself for maintenance reasons, and, further, the leakage problem is minimized. Using such a mixer, however, would necessitate a redesign of the input circuit, as the crystal outputs will then be out of phase rather than in phase, as they are now. Such input circuits have been designed for the APQ-34 and the Mk 56.

2.32 Intermediate Frequency Amplifier

The i-f amplifier consists of eight double tuned stages with an overall bandwidth of about 11 Mc/sec centered at 60 Mc/sec. The stages are identical with the exception that the grid bias gain control is applied only to the 2nd, 3rd, 4th, and 5th stages. Each stage has an average gain of about 14 db and a bandwidth of about 21Mc/sec. A typical stage is shown schematically in Fig. 2.11 and pictorially in Fig. 2.9. A mechanical drawing of the interstage transformer form is given with the receiver schematic in Fig. 2.2. Actually, in one of the receivers the transformers were mounted with their axis in line as shown in Fig. 2.9, and in the other receiver they were rotated 90° in the horizontal plane; the latter method is to be preferred, as the amplifier appeared to be more stable. The effective loading on the

* The secondary resonance frequency, f_2, is not necessarily equal to the center frequency of a low-Q double-tuned circuit. See C. P. Gadsden, op. cit.



$f = \frac{1}{2\pi\sqrt{L_b C_2}}$
Z_o=75ohms
R(=3K)
75
75 ohms
L_b
C_2
(a)
$f = \frac{f_2}{\sqrt{1 - k^2}}$
L_c
75
75 ohms
L_b
C_2
(b)
C_1
L_c
(R_1-35)=R
C_2
Mob.
75
L_z
L_b
75
(c)

Fig. 2.10

secondary consists of the input impedance of the 6AK5 tube, R_2'', which is about 7000 ohms at 60 Mc/sec, in parallel with R_2'. This results in a secondary loading, R_2, of 950 ohms. The gain of a double tuned transitionally coupled stage is

$$\text{gain} = g_m Z_{12} \tag{2.8}$$

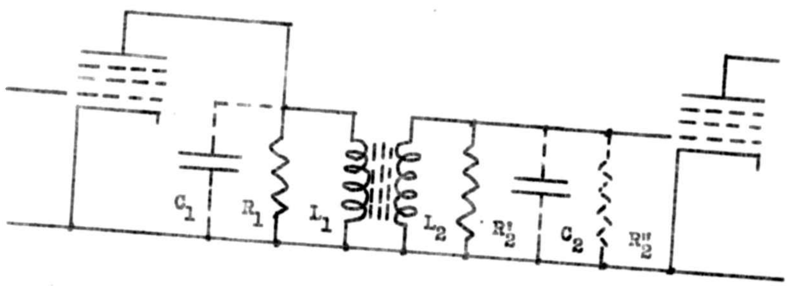
where the transfer impedance, Z_{12}, is given by,

$$Z_{12} = \sqrt{2}\, R_1 \frac{Q_2\sqrt{Q_1^2 + Q_2^2}}{(Q_1 + Q_2)^2} \sqrt{C_1/C_2} \tag{2.9}$$

and where

$$Q_1 = \omega_o R_1 C_1 \; ; \quad Q_2 = \omega_o R_2 C_2$$

In the amplifier of Fig. 2.11 the transfer impedence is 980 ohms, and assuming a JAN low limit 6AK5 tube has a g_m of 3600 µmhos, the gain is 3.53 times or 11 db. Thus the minimum overall gain that is ever likely to be observed will be on the order of 88 db.



C_1 = 4.7

R_1 = 4.3K

C_2 = 7.2 mmfd

$R_2 = \dfrac{R_2' R_2''}{R_2' + R_2''}$ = 950 ohms

R_2' = 1.1K

R_2'' = 7K

Fig. 2.11

Referring to section 2.22, equation (2.5), the rms signal voltage, E, required to just limit, was measured to be .14 volts; the rms noise voltage required at the grid of the first limiter was, therefore, found to be .12 volts. Thus the overall i-f amplifier gain must be

$$\text{req. gain} = 20 \log \frac{.12}{24 \times 10^{-6}} = 74 \text{ db}$$

It is apparent that the i-f amplifier design of the next receiver could do with one less stage. As has already been indicated, this receiver represents the first laboratory model, and at the time of its design the characteristics of the limiter were not known.

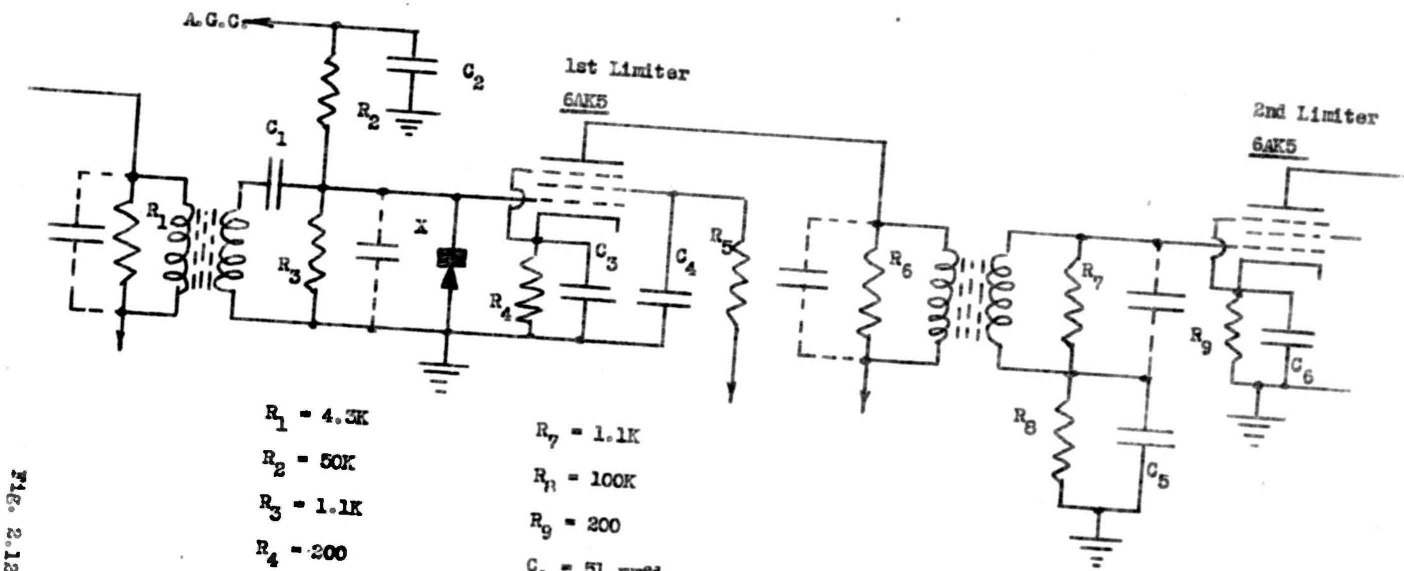
2.33 The Limiter

The most successful limiter used in this receiver is shown in Fig. 2.12. Because the design of the interstage transformer in the i-f amplifier called for a secondary loading of 1100 ohms, this resistor was used in the grid circuit of the limiter. The 51 mmfd coupling condenser makes the grid circuit time constant .056 µsec, or about 1/4 the duration of the shortest impulse passed by the i-f amplifier. Since the 1100 ohm load resistor is the same order of magnitude as the grid to cathode resistance when the grid is conducting, very little bias will be developed. The use of a germanium crystal as a diode overcomes this difficulty, as the resistance of the crystal, when conducting, is appreciably less than that of the load resistor. The same interstage transformer was used in the plate circuit as in the other i-f stages. Inasmuch as the load is too low for "plate limiting", the plate and screen voltages were made the same as the other stages. The gain of each limiter stage is about three, so that even with full plate and screen voltage, the grid of the last limiter stage is never driven to the point where the output of the second limiter decreases. Placement of the time constant of the second limiter in the "ground" side of the secondary coil was merely for convenience; its value of 50 µsec was chosen arbitrarily. When the gain of the i-f amplifier is adjusted so that noise develops 0.1 volt bias at the grid of the first limiter, the ratio of peak limit level to noise is about six or seven db.

2.34 The Discriminator

Two types of discriminators have been used successfully in these receivers. The first is the conventional Foster-Seeley type of discriminator with the primary split so as to achieve coupling on both sides of the balanced secondary. The coils were wound on a type G5 (Stockpole) iron core; details of the coil form and assembly are given in Fig. 2.13 and a photograph of the discriminator is shown in Fig. 2.14. Discriminators of this type, centered at 60 Mc/sec, have been made fairly linear over a range of about 15 Mc/sec, the bandwidth between peaks being about 20 Mc/sec.

The second type of discriminator used is shown in Fig. 2.15. This discriminator is much easier to adjust in that there is no mutual inductance and all three coils can be adjusted independently by the use of "tuning rings". Although this particular discriminator is not quite so wide as the discriminator



R_1 = 4.3K

R_2 = 50K

R_3 = 1.1K

R_4 = 200

R_5 = 400

R_6 = 4.3K

R_7 = 1.1K

R_8 = 100K

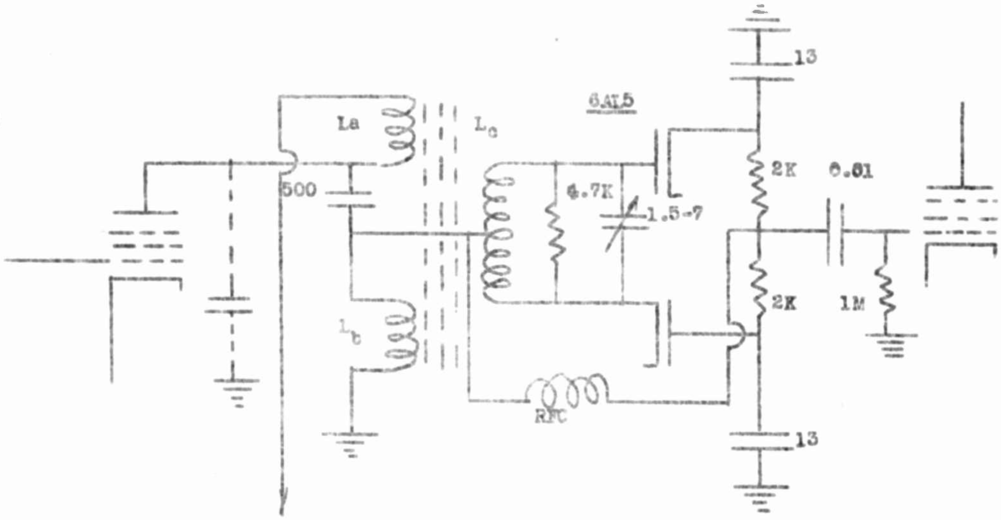
R_9 = 200

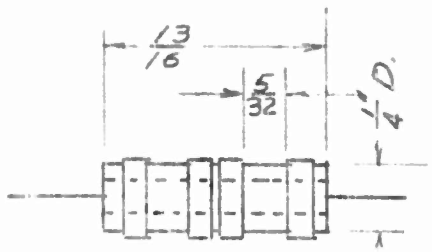
C_1 = 51 mmfd

C_2 - C_9 = 500 mmfd

X = WE #D171561 Germanium Xtal

Fig. 2.12





L_a	5T	#28 E	
L_b	5T	#28 E	
L_c	10T	#28 E	

Iron Core 0.2 " Dia.
Type G 5 (Stackpole)

Fig. 2.13

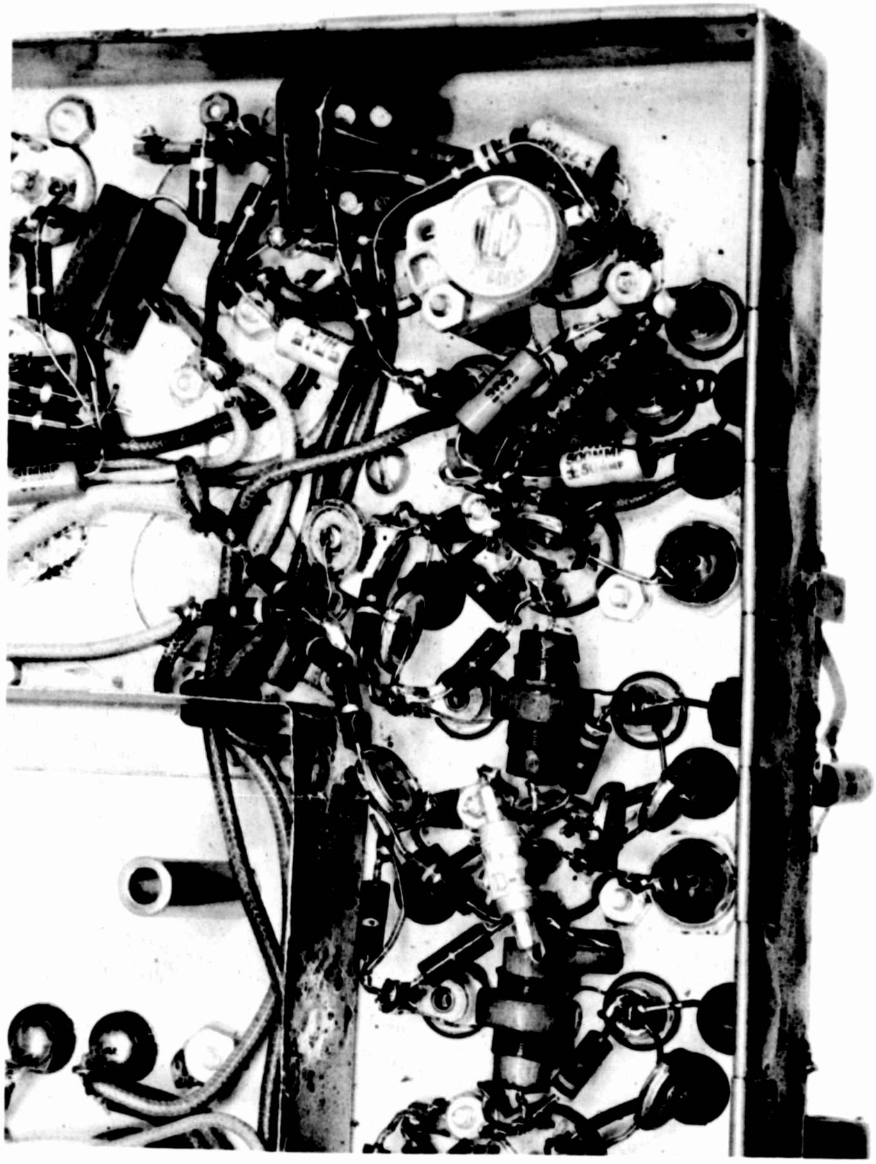
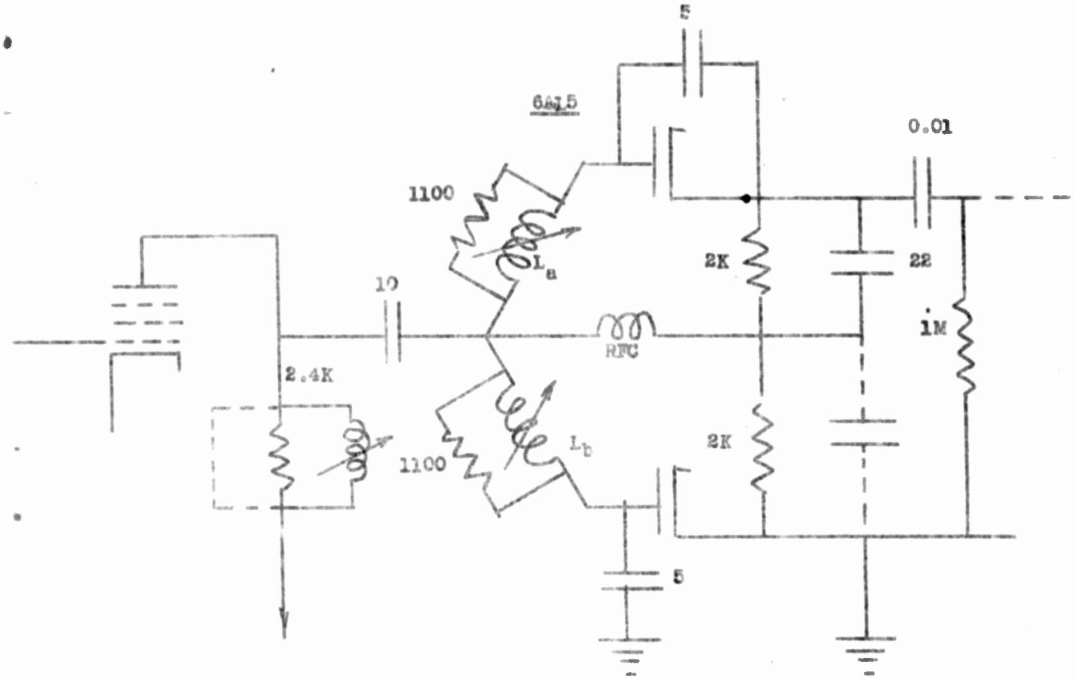


FIGURE 2-4 FM X-BAND RECEIVER-DISCRIMINATOR AND LIMITER CIRCUITS



L_a 7T #28 E 1/8 " Dia. form

L_b 14T #28 E 1/8 " Dia. form

Fig. 2.15

of the Foster-Seeley type mentioned above, there is no reason to believe that this represents an upper limit to its bandwidth. It is about 17 Mc/sec wide between peaks. Because of the relative ease in its adjustment, the second discriminator is probably to be preferred. However, it should be possible to make either discriminator "fixed tuned" in production.

In Fig. 2.16 the overall receiver passband is shown at two different signal levels. With a signal just limiting the discriminator is seen to be linear ±2% over a 6 Mc deviation. Note also that as the signal level is increased the bandwidth between the peaks of the discriminator increases. This behavior is a consequency of the presence of the limiter, and can be understood by the following explanation. In these receivers the bandwidth of the discriminator is greater than the bandwidth of the i-f amplifier. The useful discriminator bandwidth is then, of course, determined not by the actual discriminator bandwidth but by the bandwidth of the i-f amplifier. The action of the limiter, however, must also be considered. For a constant input voltage of variable frequency, the magnitude of the output voltage of the i-f amplifier will be determined by the i-f amplifier bandpass. However, the input to the discriminator will be constant and independent of the i-f amplifier bandpass as long as the signal is large enough to limit. For a given input voltage, then, the input voltage to the discriminator will fall as soon as the signal deviates far enough off resonance to be reduced by the bandpass of the i-f amplifier to the point where it no longer limits. Thus the overall receiver bandpass is seen to be a function of the input signal level, although the slope of the bandpass will remain constant.

2.35 <u>The Video Amplifier</u>

The video amplifier is of conventional design. Two 6AK5's, in parallel, are necessary to obtain the required two volts of video output across the 75 ohm line. So little gain is required of the video amplifier that it is necessary to add shunt capacitance to the coupling circuit to make the bandwidth sufficiently narrow.

2.36 <u>Automatic Gain Control</u>

The negative voltage developed at the grid of the first limiter can be used to advantage for AGC purposes. Even though the output of the limiter will be constant over a large range of input voltages, it may be advisable to use AGC to protect the later stages of the i-f amplifier from overloading. The AGC voltage is taken directly from the grid of the limiter through a high resistor, as shown in Fig. 2.12. Since it is desirable to limit on low signal levels, it is necessary to incorporate an amplitude delay of about 20 or 30 db. This delay was obtained by the use of a 6SN7, half of which was used in a diode clamp circuit and the other half as a cathode follower to bring the AGC voltage to the grids of the four controlled stages. It is shown schematically in Fig. 2.17. In Fig. 2.18 the performance of the AGC is shown. The voltage developed at the grid of the limiter was used as the ordinate rather than the actual receiver output because the output of the limitor would be constant in any event. From Fig. 2.18 we see that the delay is about 30 db.

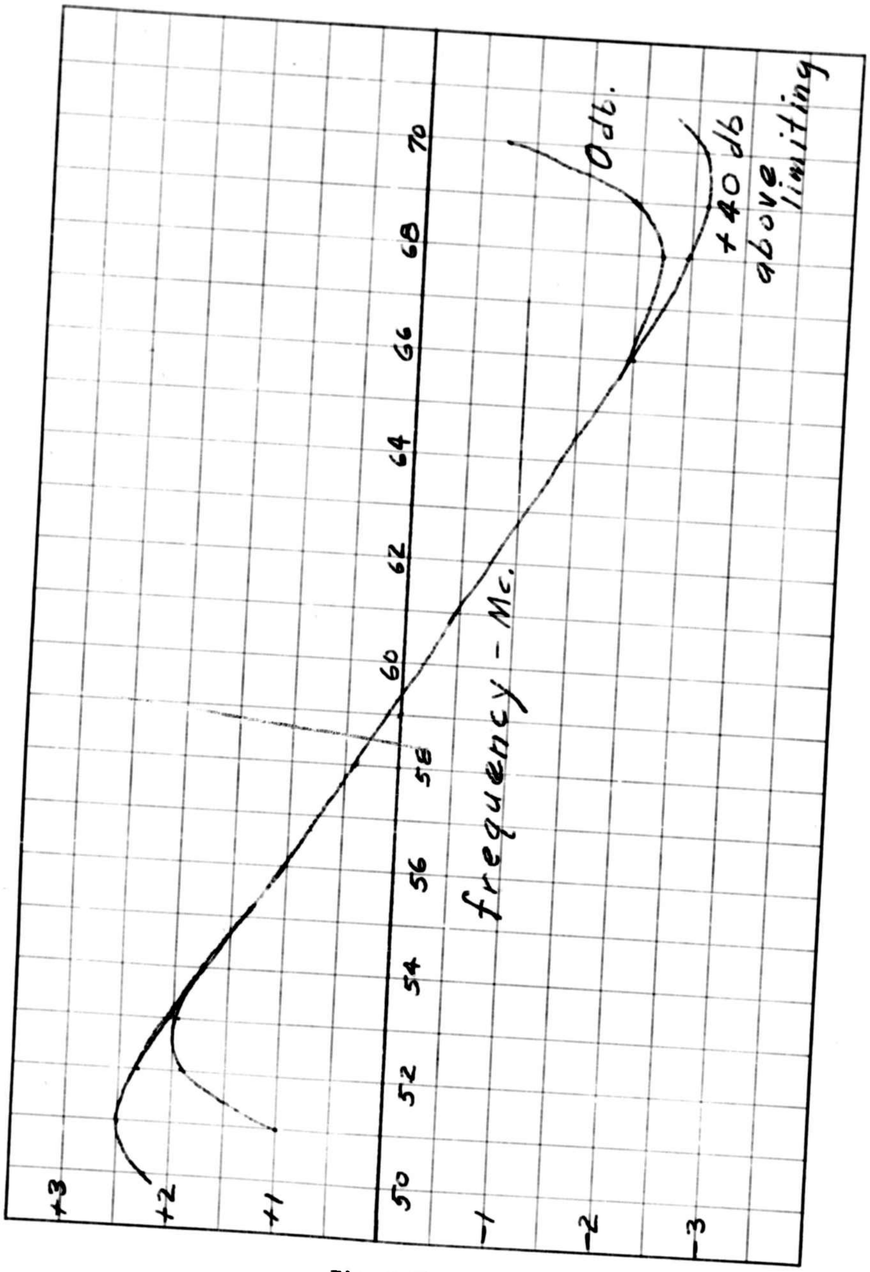

+3
+2
+1
-1
-2
-3
50
52
54
56
58
60
62
64
66
68
70
frequency - Mc.
0 db.
+40 db
above
limiting

Fig. 2.16

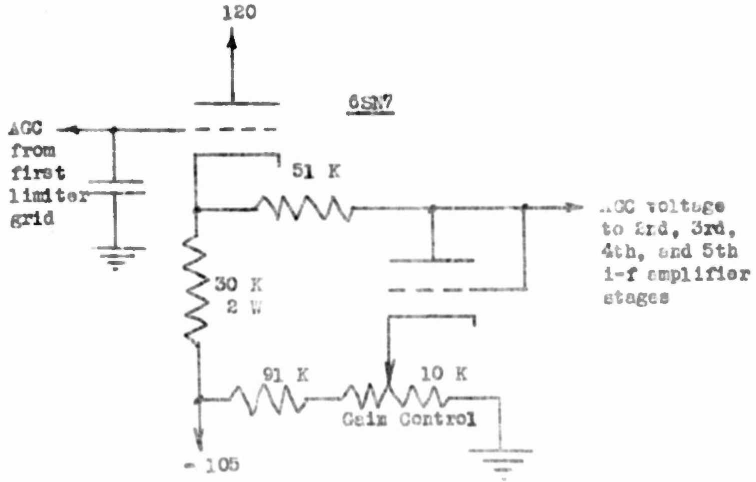


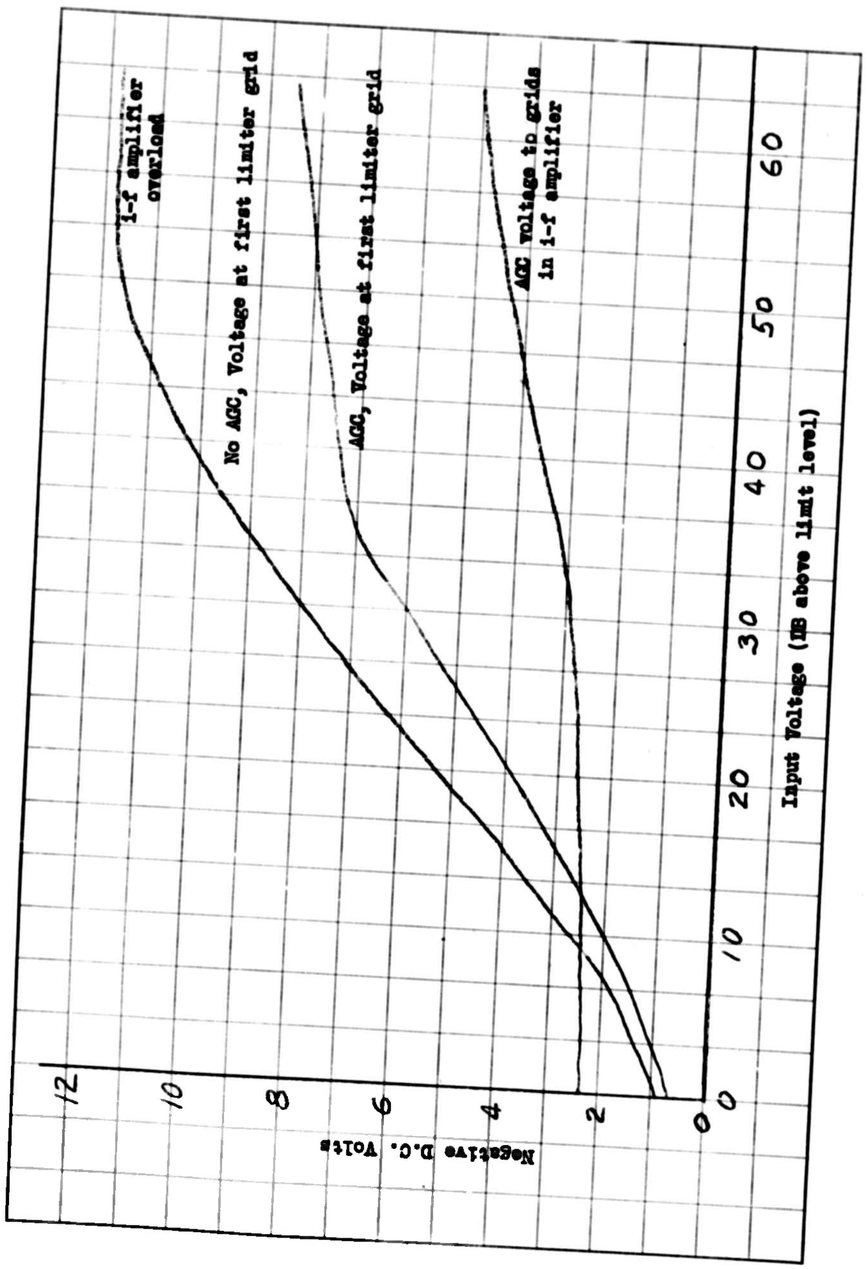
Fig. 2.17

2.37 Vacuum Tube Volt Meter

A conventional zero-centered V.T.V.M. was built into the receiver to read either the voltage developed at the grid of the first limiter or the output of the discriminator. The manual gain control is set so that thermal noise develops about .1 volt at the limiter grid. At this value, the limiter is known to operate at the "improvement threshold". The voltmeter is used in the discriminator position to facilitate tuning the receiver. The 0-1 MA meter is also used to measure crystal current.

H. Logemann, Jr.
November 29, 1945

HL:pj



i-f amplifier overload
No AGC, Voltage at first limiter grid
AGC, Voltage at first limiter grid
AGC voltage to grids in i-f amplifier
Negative D.C. Volts
Input Voltage (DB above limit level)
0
2
4
6
8
10
12
0
10
20
30
40
50
60

Fig. 2.18

ERRATA

The following paragraph should be inserted in Section 2.31 Input Circuit:-

In Fig. 2.8, L_a has been made into a bifilar coil (L'_a, L''_a of Fig. 2.7), so that the crystal current of each crystal can be measured separately. The capacitance C_1 is the total of C'_1 and C''_1, which is the capacitance of each coaxial cable to the mixer, as well as the stray capacities of the circuit. Condenser C_3 is a blocking condenser, and C_4 through C_9, with their associated RF chokes and damping resistors, make up the crystal current filter. To facilitate adjusting the circuit, a "tuning ring" (shorted turn), has been used on L_a and L_b. In production these could be eliminated. A picture of the input circuit is shown in Fig. 2.9.

Section 3.

THE CROSS-POLARIZATION MULTIPLEXER SYSTEM

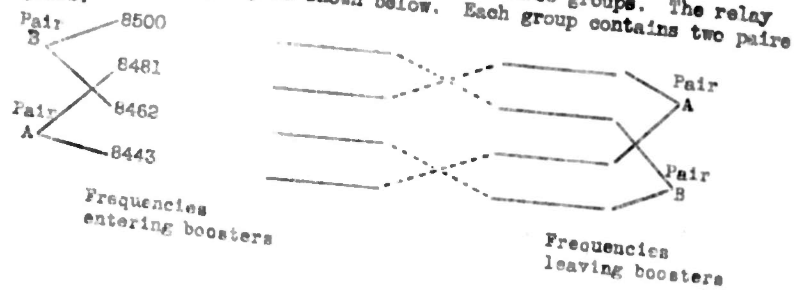
When the MEW relay scheme was proposed, no satisfactory r-f filters were available. For this reason, since it was desired to transmit only two signals from a given antenna, it was decided to employ antennae capable of transmitting two independent polarizations and to separate the two modes at the receiver by means of a mode filter. Such a scheme presumes that no cross polarization is introduced by the r-f components, antennas, filters, or by anamolous propagation affects. If this is true, then ideally this scheme permits the two frequencies to be arbitrarily close. However, since in practice some depolarization will occur, it is desirable to choose the two frequencies to be somewhat different and to rely on i-f discrimination between these two frequencies to prevent excessive interference. Thus, the role of the mode filters is to prevent 3 db power loss at the transmitter and 3 db power loss at the receiver and interaction between the generators and between the mixers.

After considering a large number of possible channel arrangements, taking into account the possibility of feedback at a booster and the frequency separation for optimum i-f discrimination, the following channel frequencies were proposed:

Group I	8500 8481 8462 8443
Group II	8424 8405 8386 8367
Group III	8348 8329 8310 8291

The channels are separated by 19 mc. The i-f frequency is 60 mc. and the frequency modulation is such that the base frequency listed here is always increased. Thus, the local oscillator must be placed on the low frequency side of the signal.

The twelve frequencies given here compose three groups. The relay scheme for a typical group is shown below. Each group contains two pairs of signals.



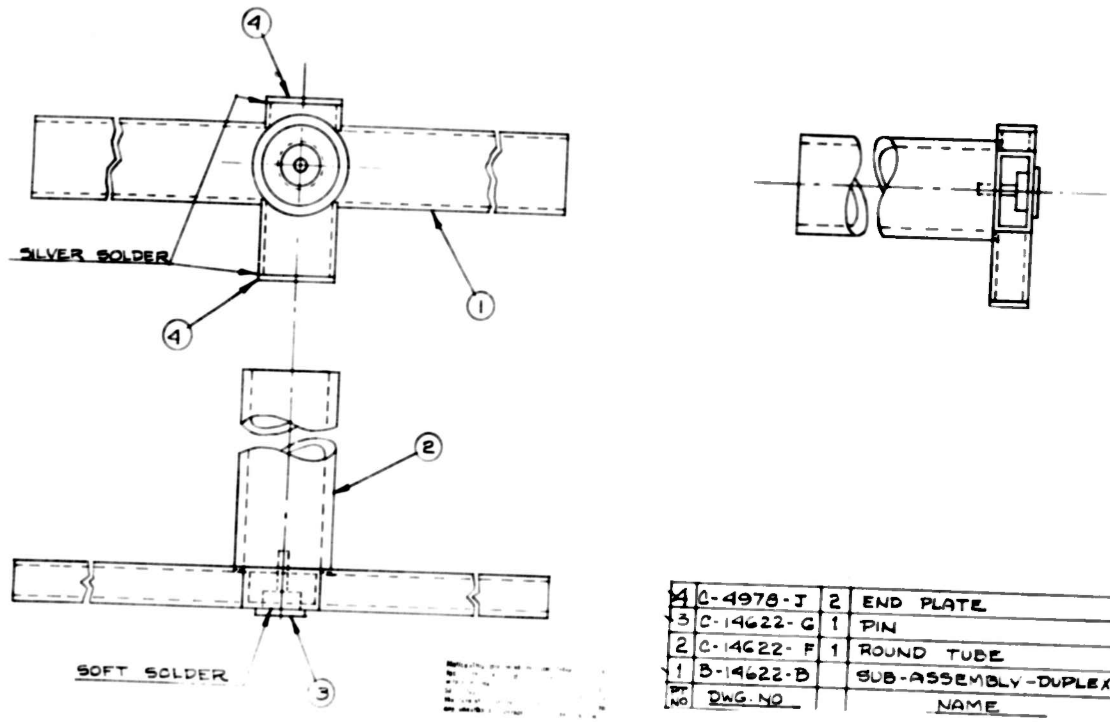
A suitable mode filter to separate two waves with orthogonal planes of polarization is shown in Fig. 3-1. This filter consists of two rectangular waveguides feeding into a circular waveguide which feeds (or separates) the waves to the antenna. No bend or other discontinuity should be made in this circular pipe if depolarization is to be avoided. Under these circumstances, cross-talk from one channel to another is down 14 db over the whole band of 200 mc.

Another type of mode filter has been tested in the laboratory and will be discussed in the Radiation Laboratory Technical Series, which cuts the cross-talk to 25 db down over the band.

No actual data on the performance of the system described above are available, but the Bell Telephone Laboratories have tested a similar scheme and found it possible to keep the depolarization to less than about 1%.

Since this proposal was made, r-f filters have been built which are suitable for building multiplexers operating on straight frequency discrimination. Such filters are to be described in the Radiation Laboratory Technical Series.

A. W. Lawson, Jr.
November 12, 1945



PT NO	DWG. NO		NAME
4	C-4978-J	2	END PLATE
3	C-14622-G	1	PIN
2	C-14622-F	1	ROUND TUBE
1	B-14622-B		SUB-ASSEMBLY-DUPLEXER

FIGURE 3-1 CROSS-POLARIZING DUPLEX FOR M.E.W. RELAY

Section 4.

ANTENNA SYSTEM

Introduction. The antenna system was not carried beyond the stage of preliminary design and some preliminary measurements. In order to clear trees, etc., at each relay station it was decided to mount the antennas on 50-ft towers. The high carrier frequency employed prohibited the use of an r-f line of this length, so that the only two alternatives were to mount the r-f parts of the transmitter and receiver at the top of the tower, or to adopt the unorthodox procedure of having the r-f equipment, including the actual antennas, at the base of the tower, with the radiated beam directed vertically upward, and to use a 45° plane reflector at the top of the tower. The latter scheme was adopted because it was felt that the convenience of having all r-f components readily available would lead to much more satisfactory maintenance in the field.

A general view of the proposed antenna arrangement is shown in Fig. 4-1. The tower illustrated in the figure is a triangular steel tower which can be assembled in a few hours. Time limitations prevented any tests as to whether this tower is sufficiently rigid for use with the narrow beam antennas described below.

Antenna System. Since it was proposed to transmit two r-f beams cross polarized, the antenna feed had to be cylindrically symmetrical. The duplexer described in Section 3, ending in 1" cylindrical guide, was terminated in a conical horn having a flare angle of 15° to a 2" aperture. The horn was arranged to feed the parabolic reflector at the base of the tower from the side in order to reduce back radiation and thus to keep the standing wave ratios in the antenna lines as low as possible. The parabolic reflector was a 5-ft square section of a 10-ft paraboloid having F = 35.8", obtained by taking 2½ feet off each end of half of the 10-ft dish. The reflector was tilted to direct the beam vertically.

At the top of the tower there was a plane reflector inclined at an angle of 45° to the vertical, which changed the vertical beam to a horizontal beam. The plane reflector was elliptical in shape so that it presented a circular aperture 4 feet in diameter when viewed from 45° to its normal.

Preliminary Antenna Measurements. The horn was found to have beam widths (to the 1/10 power points) of 92° in the H-plane and 73° in the E-plane.

The parabolic reflector and plane reflector were made of wood and metalized. Their surfaces showed some irregularities which probably led to some phase errors. The horn was adjusted for maximum gain of the system and optimum symmetry of the side lobes, after the parabolic and plane reflectors had been physically lined up by means of a surveyor's transit. Measurements of the half-power beam widths at $\lambda = 3.565$ cm led to the results given in Table 4.1.

Table 4.1

Half-Power Beam Widths of the Antenna System

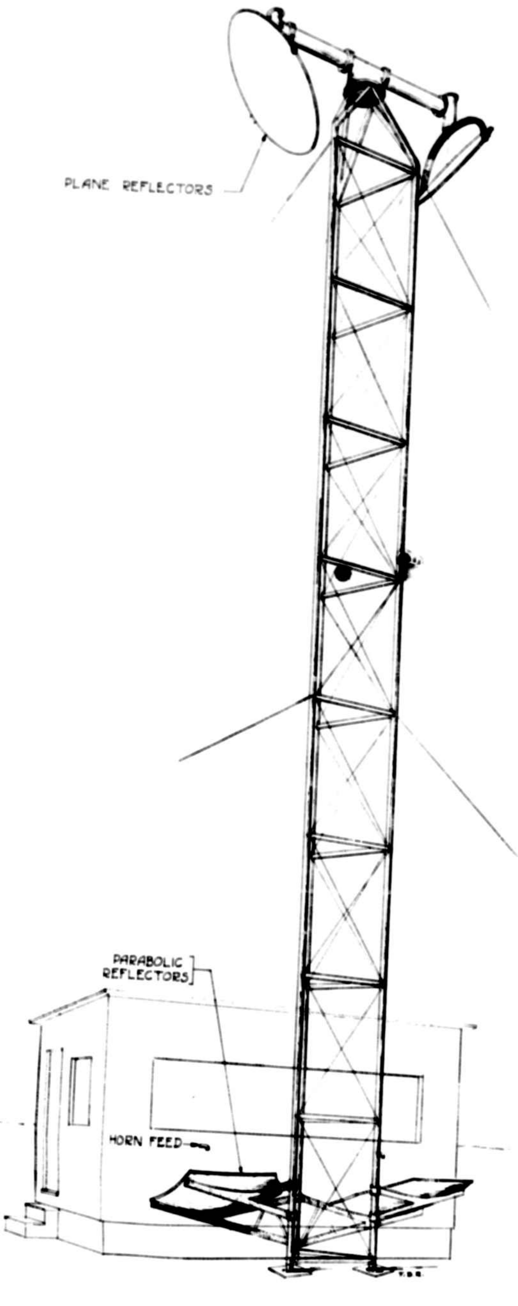
	Polarization	Plane	Beam Width
Elevation	(Vertical	E	2.8°
	(Horizontal	H	2.8
Azimuth	(Vertical	H	2.2
	(Horizontal	E	2.2

Side lobes were found to be more than 22 db down from the main lobe. The absolute gain was 5090, giving an efficiency factor of 0.442.

A 4-ft aperture at X-band should lead to approximately a 2° half-power beam width. The fact that the observed beam widths were larger is to be attributed to irregularities in the surfaces of the reflectors, and possibly also to faulty alignment between the dish and the plane reflector.

Acknowledgment. The above antenna system, excluding the steel tower, was designed and built by Group 54, and the measurements were made largely by M. Pearson. J. Whelpton helped in the design and construction of the reflectors. F. Robie of Group 101 was in charge of the tower development and prepared the sketch of the proposed arrangement shown in Fig. 4.1.

J. M. Sturtevant
November 30, 1945



REPORT 977 – FIGURE 4.1

PROPOSED ANTENNA ASSEMB'Y FOR GROUND RADAR RELAY ON 50 FT. TOWER

RFEL – C

485

A.T.I.

13830

TSDIN FORM 69 (13 NOV 47)

AUTHOR(S)		ATI- 13830
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ABSTRACT

An X-band frequency modulated relay system for video frequency signals, utilizing two specially constructed transmitters and two receivers, was tested and gave good results with one intermediate relay point. No difference could be detected between its PPI presentation and that of an unrelayed signal. The equipment of this system is discussed, and a description is given of a means by which two channels were to be transmitted and received on a single antennas.